CERN Switzerland/France, November 1998

Digital Signal Processing on Schottky Signals

Master's degree project carried out at

The European Laboratory for Particle Physics, CERN
Low level Radio Frequency group, PS/RF

for

The Technical University of Denmark, DTU,
Department of Mathematical Modelling, IMM
Section for Digital Signal Processing

by
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0.1 Preface

This report is written by me, Kristian Philip JØRGENSEN, and it is the documentation of my master's degree project.

The project is carried out at CERN in Switzerland/France for the Technical University of Denmark (DTU). The supervisor in Denmark has been Associate Professor Jan LARSEN, from the Section for Digital Processing, Department of Mathematical Modelling (IMM). The supervisor at CERN has been Flemming PEDERSEN, the section leader of the PS division low level RF section.

This report is written for people with a basic engineering background. No specific knowledge in accelerator physics, signal processing or programming, is required in advance. It has been the intention to explain all involved subjects of science, in order not to keep any potential reader from understanding the content of this report. This is basically because the environment at CERN consists of a mixture of physicists, engineers with different back grounds and techniciens, all working together with the same goal. This report is adapted to this interdisciplinary environment.

This report is written in LaTeX for better readability, mostly concerning equations. Everything is written by me, whenever there is some material taken directly from another source, it will be clearly mentioned. The figures in the report is either drawn by me in XFIG or picked from referenced material listed in the bibliography.

The notation in this report is aimed to be consistent. Whenever a reference to a book is done it is written as a number in square braces, []. All references are listed from page 126. References to tables, figures, chapters, sections etc. is referenced by the number written, when they are introduced. Only when a reference is far away from its mentioning, the page reference is written. All abbreviations are introduced with their complete list of words before their first use, but afterwards written abbreviated without additional information. The numbers used in the text are by default with a base of 10. Whenever they are different or in case it needs to be clear, their base are written as subscripts, $DEFAULT_{10}, BINARY_2, HEX_h$. Signals in continuous time domain uses t for time, discrete signals n. In the continuous frequency domain f is used for frequency and in the discrete k. The reversible Fourier transformation between these two domains are denoted $\leftrightarrow$. Whenever a [*Schaum number*] is written after an equation, it refers to [7] with *number* as equation number. Words from accelerator physics is**boldfaced** when first mentioned, to facilitate re-look up. When a program call is merged with text it is written in *italic*. When the program code is written entirely, then it is in small letters format. To facilitate the reading of this report, the use of footnotes[1] is introduced.

[1] This is a footnote

The project started the 15th of January 1998 and this report has been handed in the 16th of November 1998.

0.2 Acknowledgements

First of all, I am greatly grateful for the possibility of performing my master's degree project, for the **European Laboratory of Particle Physics**, CERN, in France/Switzerland. This is both to the **Technical University of Denmark**, DTU, which let me perform it abroad and to CERN which made it possible.

I would like to thank **Flemming PEDERSEN**, the section group leader. His dedication to this project, has been remarkable. Even nearly drowned in work, he practically always had a minute. But a minute which frequently became an hour or more. His insight in a wide range of engineering sciences is astonishing. Thank you for letting me profit from this.

Thankyou **Nick Vinod CHOHAN** for believing in this project and giving us a helping hand whenever you could. The contacts with you has been very professional and good for the progress of the project.

Another thank to my supervisor in Denmark, **Jan LARSEN**, for making a project abroad possible and for distant consultation via email. Especially the project extension, which required quite some writing back and forth.

I want thank **Maria-Elena ANGOLETTA**, for the hours we have worked with the system. You have a high level of knowledge in programming and it has been interesting to solve the numerous system malfunctions, with you.

A personal thanks to, **Silvia GRAU** for understanding and support during stressful periods of the project. As well to **all of my friends** that I have made around Pays de Gex and the city of Geneva, your company has been very nourishing.

0.3 abstract

This report covers the development of an embedded VME crate data acquisition and processing system. The system is meant for processing of detected transverse and longitudinal beam signals, from the AD synchrotron at CERN. The sampled beam signal data rate is close to 2 giga samples per second, so state of the art processing hardware combined with signal processing principles is used. A frequency band of interest is downconverted by a digital receiver (Pentek 6510) and a digital signal processor (TMS320C40) processes data, in an real-time fashion. The code in the digital signal processor is optimised for speed, in order to meet real-time processing constraints. This is done by writing a mixture of assembly and C code.

Chapter 1

Introduction

This chapter starts out with an introduction of the project scope. Following is an introduction to the environment at CERN. Some of the activities and key features of the organism is mentioned. The group, in which this project has been carried out, is introduced in the section following.

1.1 Scope of project

The scope of the project, can be boiled down to a short sentence which coincides with the project title.

"Digital Signal Processing on Schottky Signals"

This includes quite a lot of tasks. First of all there is the understanding of the signal source which originates from physical processes. This understanding is essential, in order to implement the right processing tasks. Then there is detection of signals transforming the physical quantities into electronic signals. This is done elsewhere, thus only a peripheral knowledge is required. The main part of the project is the acquisition and processing of these electric signals. This comprises everything concerning development of such a system. The main part of this project can thus be defined more specific as following.

> **Design of an embedded VME-crate-based real-time acquisition and processing system for power spectral density analysis of signals from four independent parallel sources at high data rates.**

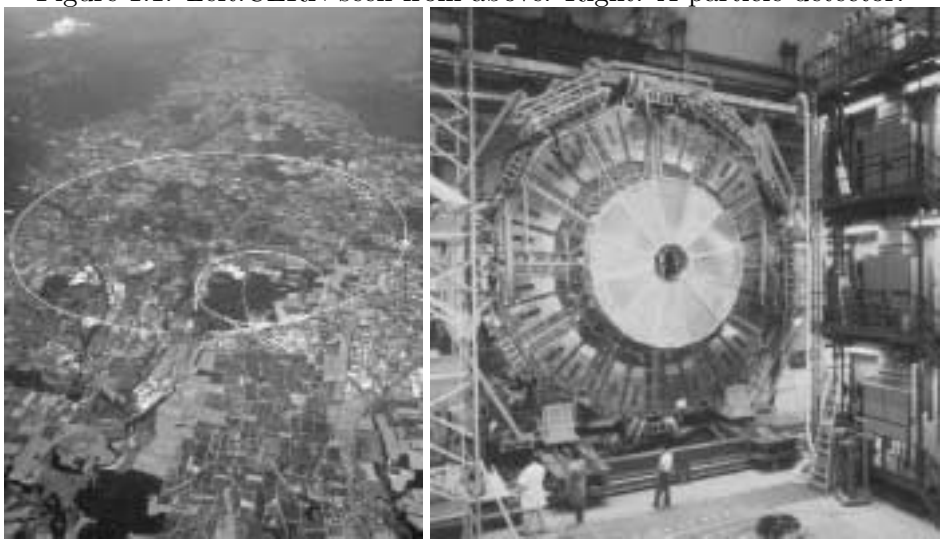
1.2 Project environment

1.2.1 CERN

CERN is the European Laboratory for Particle Physics originally an abbreviation for the french, Centre European de la Recherche Nuclaire. It is the worlds biggest particle physics centre founded in 1954. The 12 european member states all financially contribute to the research and the annual budget is around 600 million US dollars. The member states contribute with a certain percentage and this should correspond to the percentage of employees from the member states, taking part in the work at CERN. There are around 3000 employees payed directly by CERN and 6500 scientists using the facilities at CERN.

The overall goal of CERN is research of particle physic and for this is used rings for accelerating, decelerating and storing particles for different purposes. The largest one of those rings are 27 kilometres (LEP[1]) on both swiss and french territory, see figure 1.1. It is situated 100 meters beneath the ground surface and the 4 detectors are of the size of four storey houses. Particles travelling along this 27 km. circular accelerator travels near the speed of light, this means that they make around 11.000 rounds per second. This and other tasks require highly specific equipment which only finds its use at CERN. That is why a lot of development is done at these premises. As an example, CERN has a magnet in an accelerator which weighs more than the Eiffel tower in Paris.

Figure 1.1: Left:CERN seen from above. Right: A particle detector.



In some circular accelerators two beams collide in particle detectors.

[1]Large Electron-Positron collider

This is to discover what mass is made of. The particles are broken into their components and their trajectory are measured by the particle detectors. One of the quests of this decade is the search of the Higgs particle, which should exist, but this is never verified. The resolution of the particle detection rises with the momentum of the particles. That is why the bigger the accelerator the better. The particle detectors are high technology constructions. They measure among others the trajectory of particle components, from which their mass and polarity can be calculated. The data acquisition from some particle detectors are astronomical. Some of these particle detectors have data rates which would be equivalent to the event that every person on earth, would make 10 telephone calls, simultaneously. The task of making this possible employs a lot of engineers. Moreover the data from these experiments should be accessible for scientists on an on-line basis.

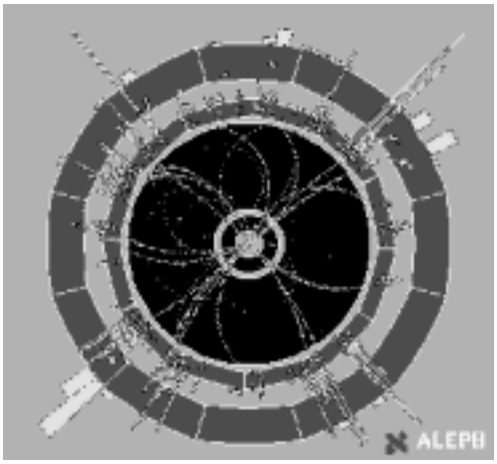


Figure 1.2: Beam collision

Other experiments with particles are done at rest. This means that special particles are created at high energies and then decelerated in a circular decelerator. They are then let out of the ring and held in something called a Penning trap.

Even though the purpose of CERNs activities are particle research there is development in a lot of other fields. The World Wide Web, for example, is developed at CERN for the purpose that physicists could share data from experiments instantaneous, now matter where they were.

As CERN is a research centre there is a lot of publishing and lectures on recent discoveries and new experiments. These are in all areas of CERN activities, but with a majority in the area of physics. CERN is, as well, a place chosen for a lot for conferences. These conferences vary from basic accelerator theory to highly specific areas of physics. Conferences are mostly free for people working at CERN, but normally open for people

from outside CERN.

A good popular explanation of CERN activities and particle accelerators is explained at the CERN homepage at:

http://www.cern.ch

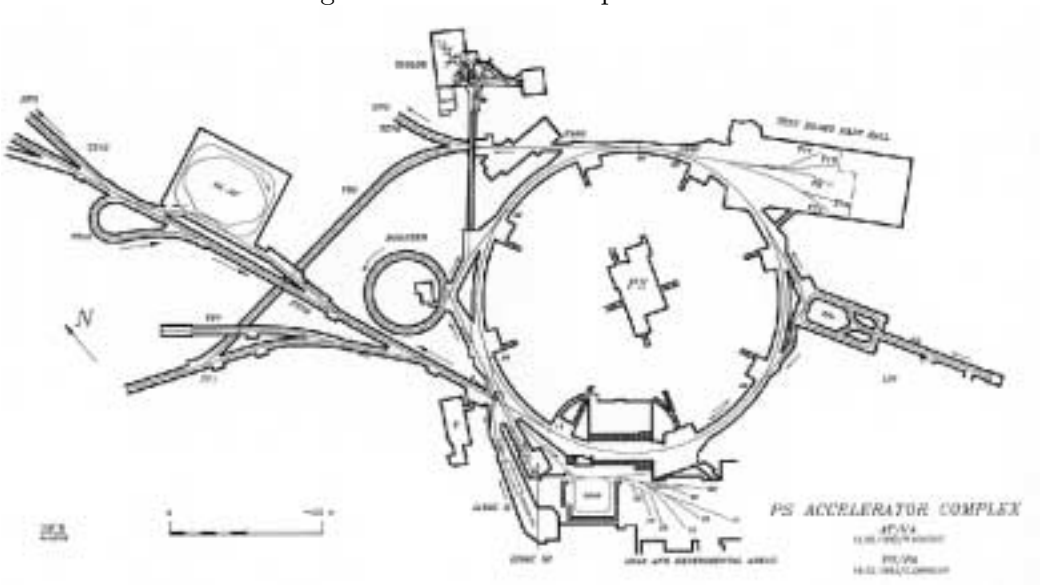
1.2.2 Introduction to the PS/RF

The PS/RF group is the radio frequency group in the PS[2] department. The RF group is responsible for more than 50 different RF systems in the PS complex, see figure 1.3. The group is further divided into three minor groups which are:

- Low Level RF
- High Power RF Circular Machines
- High Power RF Linacs and RFQ

The frequencies worked with in the RF group ranges from 0.6 MHz to 3 GHz and the powers from milliwatts to megawatts. This project is done in the Low Level RF group. The project doesn't touch the high region of radio frequencies, but the equipment developed in this project tests that other modules that do, is working properly.

Figure 1.3: The PS complex



[2]Positron Synchrotron

One of the challenges of the group is the AD project of which this project is a part. The AD project involves RF "gymnastics" in order to control the beam with less resources. The AD project is due to finish around April 1999.

Chapter 2

Accelerator physics

Accelerator physics is a complicated matter for engineers, not having to do with physics normally. This project, however, needs to introduce some basic properties from the accelerator physics, in order to introduce the problem which this project is supposed to solve. There is quite a lot of specific terms and properties which are introduced in this section. The introduction is brief and can be skipped by persons familiar with general accelerator physics. A complete description of the processes is outside the scope of this report, but for further details see [3, 5, 4, 6]. The important terms introduced is boldfaced when explained,to facilitate re-look-up.

2.1 Reason for the AD project

The AD[1] project is meant to replace the older method of producing and decelerating antiprotons for antiproton experiments at rest. This new configuration only involves a single synchrotron, the AD, for the collection, cooling and deceleration of antiprotons. Previously, 4 synchrotrons where used to fulfil the same function, namely the AC[2], the AA[3], the PS[4] and the LEAR[5]. This is a restructuring of the AC, to a new configuration, AD, which will lead to a more economical way of producing antiprotons. In addition it will be a lot faster to produce them. The economical factor lies in the fact that the former installations needed large amounts of electrical power, to produce the magnetic fields for the previously used storage rings AC and AA. The electrical power was an expense of 16-20 Million Swiss francs[6] per year. The new AD synchrotron only operates at full field with

[1] Antiproton Decelerator
[2] Antiproton Collider
[3] Antiproton Accumulator
[4] Proton Synchrotron
[5] Low Energy Antiproton Ring
[6] Which is approximately equal to 13-17 Million US dollars

a duty cycle of less than 16%. These fields are proportional to the current and the power rises with the square of the current, so by minimising the period with high currents, a lot is gained. When decreasing the areas where magnetic fields are needed, the total power consumption is reduced remarkably, as well.

Another important reason was that a lot of resources had to be freed for another CERN project, LHC[7], which has another history of its own.

2.2 AD lattice

The antiproton decelerator (AD) is a synchrotron, which consist of a vacuum chamber where the particles are guided by specific cells, with magnetic and electrical fields. These fields insures that all particles, more or less, follows a well defined trajectory around the installation. These cells are in order of description, bending, acceleration and focalisation. The configuration of such a circular machine is called a **lattice**. The lattice is sketched at figure 2.1, which gives an idea of the distribution of cells along the 182.43 meter long lattice.

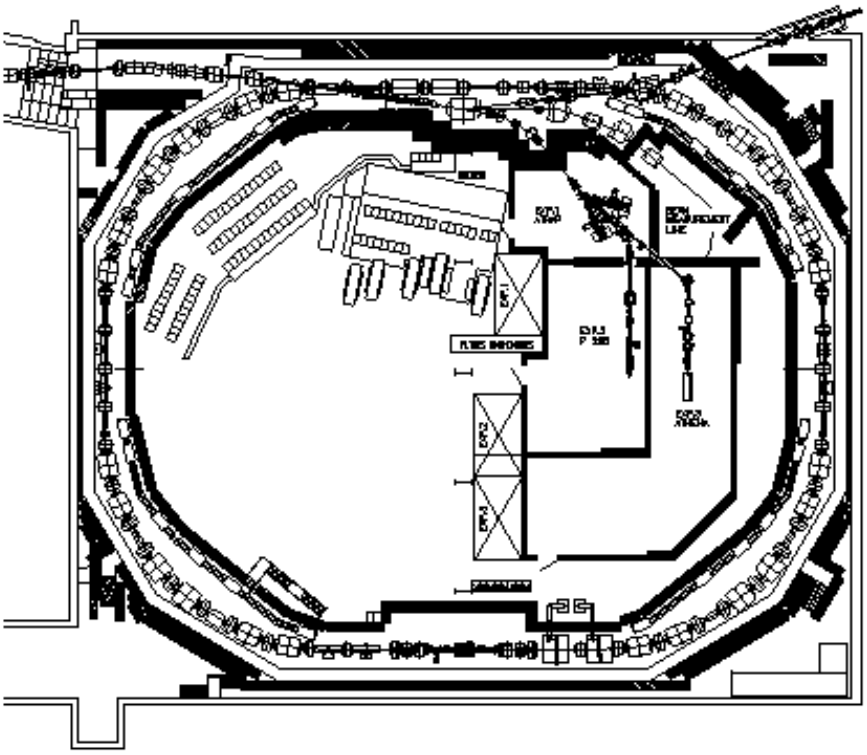


Figure 2.1: The AD lattice

[7]Large Hadron Collider

Additional information about the synchrotron structure can be found at [5] and [3]. Specific information about the AD project is described in [4].

Bunch

The particles can circulate in two modes **bunched or unbunced**. The unbunched is with random positions along the trajectory, whereas bunched is when the particles are kept together by horizontal electric forces.

Synchronous particle

The **synchronous particle** is an abstract reference particle that perfectly follows the designed trajectory. The synchronous particle has the same trajectory in every turn and the trajectory coincide with what is called the **closed orbit**. If there are no coherent oscillations among the particles travelling together in a bunch, then the synchronous particle coincide with the centre of massof the bunch.

Betatron/Synchrotron oscillations

Practically no particles follows the closed orbit perfectly, there is always some oscillation about this orbit, of minor or larger amplitude. The transversal decomposition of such a 3 dimensional motion is called the **betatron oscillation**. A longitudinal decomposition of the oscillation is called the **synchrotron oscillation**. The betatron oscillation is also denoted the **tune** and it is given as $Q = \frac{f_{betatron}}{f_{rev}}$. The respective cells affect either the betatron or the synchrotron oscillation. No cell affects both, so these two decompositions can be considered as being uncorrelated.

These two quantities are of great importance to the project, please make a note of them.

Vacuum chamber

The beam travels inside a **vacuum chamber**, which is held at a very low pressure to avoid the particles from colliding with gas molecules. In the former AC configuration the total pressure in the vacuum chamber was around 8 pbar., This is equal to a total molecule density of less than 1015 [Molecules/m3]. The vacuum conditions for the AD configuration has to be improved 20 times.

Energy

The particles have a total energy that can be divided into two contributions, the kinetic, T, and the energy at rest, E_0, $E = T + E_0$. The energy at rest

is calculated as $E_0 = m_0c^2$, where m_0 is the mass at rest and c the speed of light. As the velocity of the particle rises, it gains more kinetic energy. The contribution from the kinetic energy is, $T = (\gamma - 1)E_0$. γ is the relativistic factor given as

$$\gamma = \frac{m}{m_0} = \frac{1}{\sqrt{1 - \frac{v^2}{c^2}}}$$

- where v is the velocity of the particles and m the relativistic mass. It can be interpreted as , the factor that the mass of a particle, gains by having a velocity. In classical mechanics the formulas are similar , only the velocities are very small compared to the speed of light, so the γ factor is close to 1.

Approaching the speed of light makes γ approach infinity and the kinetic energy becomes significant compared to the rest energy. Close to the speed of light the velocity of a particle changes very little, as result of rise in total energy. The rise in energy level is stored as extra mass in stead. This is why the particles are never accelerated to the speed of light, this would require disposal of an infinite energy source. The quantity that gives the ratio between the velocity and the speed of light, is called β, $\beta = \frac{v}{c}$.

The momentum of a particle is, $p = mv = \gamma m_0 v$. It is measured in $[eV/c]$, thus multiplied by c we have pc with the dimension of energy. In the normal SI[8] system, energy has the dimension Joule, $[J]$, but in accelerator physics the unit electron volts, $[eV]$ is used, as it is closer to the related motion. One electron volt is defined as the energy an electron gains, by accelerating through a electrical potential of one volt. Likewise, we can calculate the mass as being proportional to energy by multiplying with c^2. As an example, a proton (and antiproton) has the mass

$$m_0 = 1.670\ 10^{-27}\,[kg] = 938\,[MeV/c2] \Rightarrow m_0c^2 = 938\,[MeV]$$

This way a comparison is possible and at certain velocities $v \to c$ the contribution from energy at rest can be neglected.

The formulas for relativistic calculations mentioned above, can be summed up in table 2.1. These formulas are used in later sections, without reference to the table 2.1.

2.3 Cells in the lattice

Bending Magnets

Bending is the action that accelerates the particles horizontally, towards the centre, so that they describe a circular loop. By using uniform magnetic fields, normal to the velocity, the particles are subjected to a force, that

[8]Standard Internationale

Table 2.1: Relativistic formulas

Total Energy	$E = T + E_0 = \gamma E_0$	$E \approx$ pc for velocities near c
Rest Energy	$E_0 = m_0 c^2$,	$E_0 = 938[MeV/c]$ for a proton
Kinetic Energy	$T = (\gamma - 1)E0$	
Momentum	$p = \gamma m_0 v$	
Relativistic Factors	$\gamma = \frac{m}{m_0} = \frac{E}{E_0}$ $\beta = \frac{v}{c}$	$1 < \gamma < \infty$ $0 < \beta < 1$

makes them accelerate towards the centre. The momentum derivative is described by Lorenz as:

$$\frac{dp}{dt} = e(\overline{E} + \overline{v} \times \overline{B}) = \overline{F}$$

- where e is the charge, F is the force vector.

Having a vertical uniform magnetic field and a trajectory completely in the horizontal plane. Using a system of coordinates with s as a tangential vector along the curvature, x as horizontal component and z as vertical. The equation can be reduced and split into the 3 components:

$$(p_x, p_s, p_z) = e(v_s B_z, 0, 0)$$

,which describes a circle trajectory with radius,

$$\rho_x = \frac{p_s}{|e|B_z} = \frac{\gamma m v_s}{|e|B_z} \; [m].$$

The uniform magnetic field is obtained by a magnetic dipole.

The momentum is proportional to the bending radius, ρ , in the magnetic field, B. The strength of the magnetic field is inverse proportional to this radius. In a synchrotron the magnetic field and the momentum is synchronised, $|p| \propto |B|$, thus resulting in conservation of the bending radius.

This principle has a great advantage. It limits the trajectory of the beam to be a closed loop, in stead of a spiral as in cyclotrons.

In the AD lattice there are 24 of those bending magnets, each of them taking care of 15° ($\frac{360°}{24}$) of the bending angle.

RF cavities

The velocity of the beam is controlled by **RF**[9] **cavities**. They have two functions, to accelerate/decelerate a bunch and to divide and keep the beam

[9]Radio Frequency

divided in bunches. This cell has an impact on the synchrotron motion of the beam. A simple cavity is a plate with a hole for passages of particles. Between the plates there is a voltage resulting in an electrical field between the plates. With alternating voltages, at radio frequencies, connected to the plates, it is possible to focus the particles longitudinal around the synchronous particle. The voltage ($V = V_{plate1} - V_{plate2}$), between the plates, are sinusoidal and varies with time as, $V(t) = V_0 sin(2\pi f_{RF} t)$. The radio frequency, $f_{RF} = \frac{hv_s}{L_{closed\ orbit}}$ is a multiple of the revolution frequency $f_{RF} = hf_{revolution}$, where h is the integer number of bunches along the closed orbit, $L_{closed\ orbit}$ the length of the closed orbit and v_s the velocity of the synchronous particle.

For the AD synchrotron, the length is 182.43 meters. There is 1 bunch (h=1) and to start with, the particles are travelling with the speed that is 96.72% of the speed of light ($\gamma = 0.9672$).

Thus this system needs an RF frequency of:

$$f_{RF} = \frac{hv_s}{L} = \frac{1 \times 0.9672 \times 2.9979\ 10^8}{182.43} = 1.59\ [MHz]$$

At each passage of the RF cavity, the particles are either accelerated, decelerated or neither. The change in longitudinal kinetic energy is given as:

$$\Delta E_s = qV_0 sin(\phi_s)$$

- where ϕ_s is the phase of the arriving synchronous particle, in respect to the RF frequency, and ΔE_s is the synchronous energy gain per turn. This gain corresponds exactly to the change of the strength, in the magnetic field, B.

There is a small dispersion of phases in the bunch of particles, so not every particle is arriving with this phase, nor the same energy. For a non-synchronous particle, with an energy difference in respect to the synchronous particle of ΔE_n, the energy difference after passage becomes:

$$\Delta E_{n+1} = \Delta E_n + |q|V_0(sin\phi_n - sin\phi_s)$$

so if the phase, ϕ_n is bigger than ϕ_s (but still smaller than $\pi - \phi_s$) ,then the energy difference, compared to the synchronous particle , will grow. But in a stable case, the phase will decrease within the next iteration. The next iteration will have another phase and thus another change in the energy level. The result is a trajectory about the synchronous phase, ϕ_s, which is closed for phases, ϕ_n , that doesn't exceed $\pi - \phi_s$ for zero value energies, ΔE_n, (see figure 2.2).

For small variations the trajectories are close to circles, for larger amplitudes they take form as fish-like trajectories. The largest possible trajectory that insures stable longitudinal synchrotron oscillation is called the **separatrix or bucket**. A particle moving along coordinates $(\Delta E_n, \phi_n)$ outside

this closed trajectory will follow an open trajectory and is unstable.
If the differences in phase are too significant, then the rise in energy level will be big. Particles may then not be bent enough, by the bending magnets, and they will follow a trajectory along a slightly larger circumference. Then they will follow a longer route and they wont be able to catch up with the synchronous particle, in spite of their rise in total energy. Such unstable particles will grow in both phase and energy difference. Eventually they will be lost, due to geometrical limitations of the vacuum chamber. These are the particles that follows the open trajectories outside the separatrix.

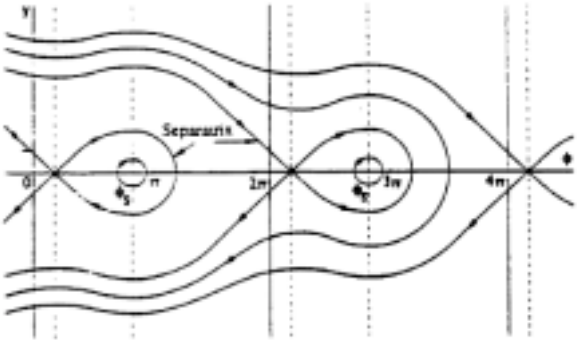


Figure 2.2: Synchrotron motion in phase plane

Quadrupoles

If ideal fields were available, there would be no need of focusing the beam onto its orbit. But due to imperfections, in the fields, the beam is not kept together automatically. Small variations in the particles grows and if no step towards dispersion were taken, it would not be possible to conserve a beam, in a storage ring, for very long time.

A weak focusing force can be implemented, by adding a gradient to the guiding magnetic field. This way a vertical focusing could be obtained. In the bending magnets a gradient could be added to insure horizontal focusing. This is possible but large volumes of magnetic fields are required.

A better focusing mechanism has thus been invented. By adding what is called an **alternating gradient or a strong focusing force**, it is possible to obtain small amplitudes of the betatron oscillations. Thus a smaller beam dimension is needed, which also makes it possible to raise the magnetic field along the orbit. Thereby, both higher energy and higher beam densities, can be obtained.

This force comes from focusing magnets called **quadrupoles**. They consist of four poles, constructed in a way that the centre of the magnet,

has no magnetic field and the field is rising linearly, with the displacement from the centre (see figure 2.3).

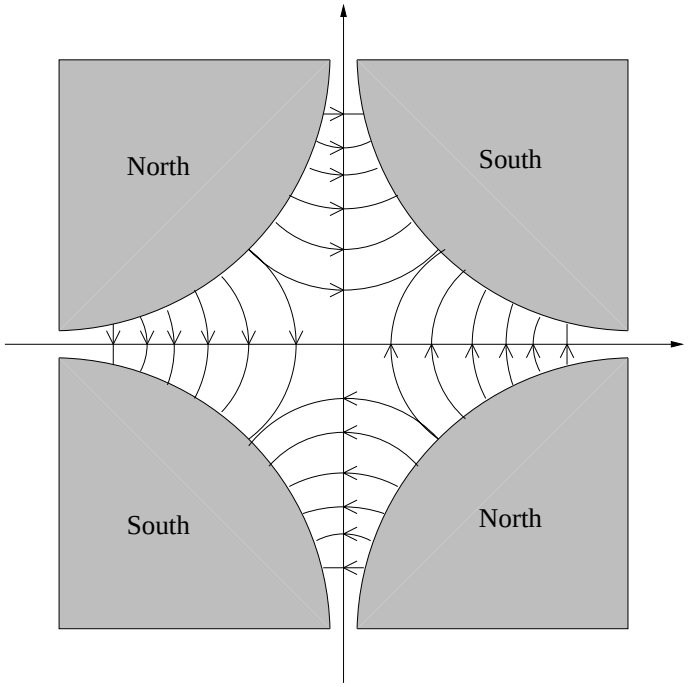


Figure 2.3: Quadrupole focusing horizontally for antiprotons going into the paper

There are two types of quadrupoles, only difference is the orientation of the poles. They are respectively flipped 90 degree clock or counter-wise to one another. This results in gradient fields that are successively focusing along one axis and in the next focusing along the other, 90 degree flipped axis.

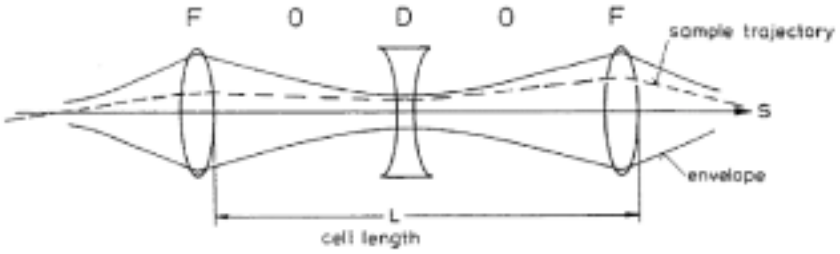


Figure 2.4: FODO cell

Looking at only one component , say the horizontal axis, a particle meets a focusing magnet (F) that pulls the particle towards the centre of

the quadrupole. During the subsequent piece of orbit, there is no focusing forces/effects involved(O). Then the particle meets a focusing force in the vertical component, but due to the flipped structure it is now defocused in the horizontal plane(D). The defocusing is proportional to the displacement from the centre of the quadrupole and as the beam is smaller horizontally at this location, defocusing will not be as strong as focusing. The lattice is then build up of a system of these quadrupoles in a, so called, FODO pattern. Such a pattern has successively changing magnet polarities. The AD lattice has 28 such FODO cells. 56 times during rotation the angle, with respect to the closed orbit, of the particle is changed. This eventually results in transversal oscillations around the orbit.

The motion around the transversal plane is somewhat discrete, as the cells result in almost instantaneous changes and the change of displacement is linear with time. The normalised frequency of these transverse betatron oscillations are denoted Q. This is the frequency in respect to the revolution frequency,($Q = \frac{f_{betatron}}{f_{rev}}$). It is also sometimes denoted the **tune** of the beam. In the AD lattice, the betatron oscillations has Q values in the horizontal plane of $Q_h = 5.39$ and in the vertical plane $Q_v = 5.37$. So these particles with a Q in the neighbourhood of 5.3 needs approximately 11 ($\approx \frac{56}{5.3}$) passages of quadrupoles, to perform one single oscillation. The analogy to lenses, as shown on figure 2.4, are thus an example fairly simplified. The displacement is more like on figure 2.5 , here there is 76 changes of derivative, which is similar to sampling the position after each quadrupole. This is a typical path for the particles.

The analysis of cells affecting the beam trajectory, is however called the optics because of the similarities with lenses.

Kicker magnets

The **kicker magnets** are used to inject or eject a beam from the lattice. This is a dipole magnet, with steep rise and fall time, insuring rapid change of beam trajectory. The reason for having this cell, is that the bending magnets has strengths of fields that do not easily enable rapid changes. So the magnetic fields of the bending magnets are conserved and in stead the beam, or part of it, is guided passed the bending magnets.

Stochastic cooling

Stochastic cooling is performed when the emittance has to be brought down. This is only one of the many ways of cooling the beam (electron cooling, laser cooling etc.). It is done by measuring a fraction of particles, n, of the total amount of particles, N. As the number measured is much less than the total number, $n << N$, it is more likely that the particles

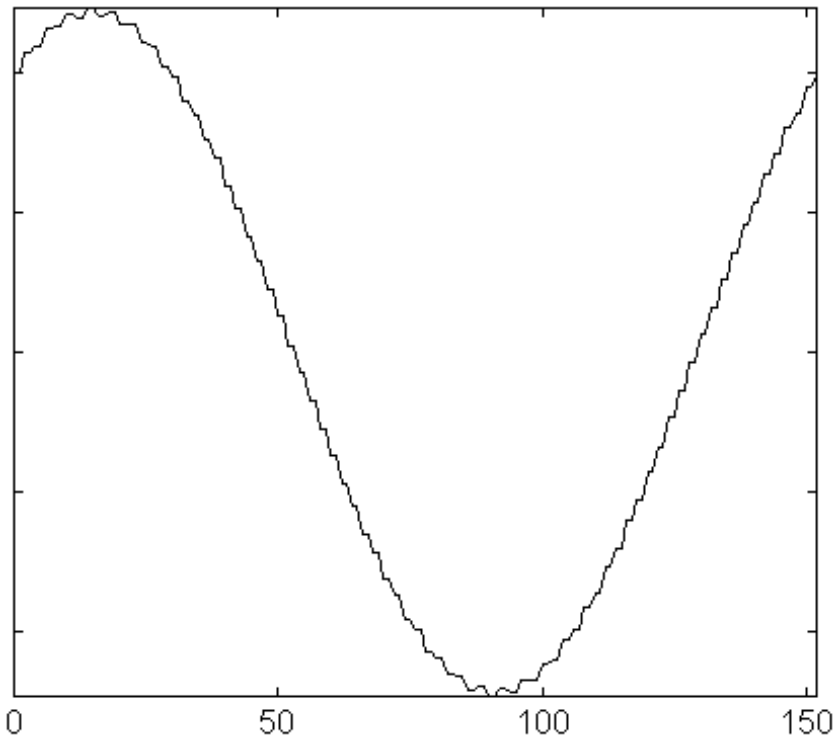


Figure 2.5: Betatron oscillation for a single particle

have a local displacement. Whereas for larger number of particles the measurement would approach the average of zero displacement. Kicking towards the orbit can eventually eliminate the mean displacement, of these n particles. The total emitance thus decreases.

2.4 Beam dynamics

Creating a moving reference frame, for the particle(s), with the closed orbit as origin, enables an easier description of the motion of the particle(s) in rotation.

The motion of particles is close to Hamiltonian motion. This is similar to the motion of a pendulum, where the Hamiltonian energy is stored as either potential or kinetic energy. As the particles aren't subjected to resistance on their travel, ideally, they will perform undamped Hamiltonian oscillations.

The motion is thus Hamiltonian oscillations along three axes, longitudinal, transverse horizontal and transverse vertical. The resulting motion is some bizarre 3 dimensional path about the synchronous particle. Decomposing and treating the motion in these three planes is possible, as the

lattice is build so that motion in one plane, does not have influence on the others.

In the longitudinal plane the delay of a particle along the closed orbit, in respect to the synchronous particle, is denoted τ_i. The particles will perform a sinusoidal oscillations about the synchronous particles. The delay will, ideally, follow the equation:

$$\tau_i(t) = \hat{\tau}_i sin(\Omega_s t + \Psi_i)$$

, which follows the synchronous particle in mean, but has sinusoidal fluctuations about it. At a flash in time, the particle has thus a delay in respect to the synchronous particle, but also a different level of kinetic energy. That is, the particle might be behind or before, but accelerates so it eventually approaches the synchronous one (stable state). The resulting motion is best shown in a **phaseplane**. The difference in phase from the synchronous particle as first axis and difference in energy level, ΔE, as second. In this plane the motion will be a ellipsoidal motion about the origin. The origin is a fixed reference, being the synchronous particle.

In the transverse horizontal and transverse vertical plane, the same axes are used. First axis is the displacement from the closed orbit and second the derivative of the displacement, hence velocity. As before we decompose the movement in a horizontal and a vertical plane. The oscillation, about the orbit, will in this system, as well, be described as an ellipsoidal motion about the origin in both these two phaseplanes.

For the same example as used prior to this section, a plot of 76 passages of quadrupoles is shown on the figure 2.6.

The solid line is the complete movement of particles, whereas the dotted line is drawn from discrete samples after each FODO cell.

2.4.1 Motion contributions

The cause of these movements in the three planes, can be decomposed into several linear contributions. The **bending**, the **acceleration**, the **focalisation** and the **non-active straight sections** of the lattice.

The action of **acceleration** has an effect on the longitudinal phaseplane. By passage of the accelerating cell, the particle increases its kinetic energy (upwards in phaseplane) almost instantly. Within the section between the next accelerating cell, it increases the phase, $\phi(t)$. This will create a somewhat bizarre path about the origin, made of sections of straight lines in the longitudinal phaseplane. These lines are quite small, compared to the circumference of the trajectory around the phaseplane.

The action of **focalisation** is likewise described by an almost instantaneous change of angle, in respect to the closed orbit. In between the

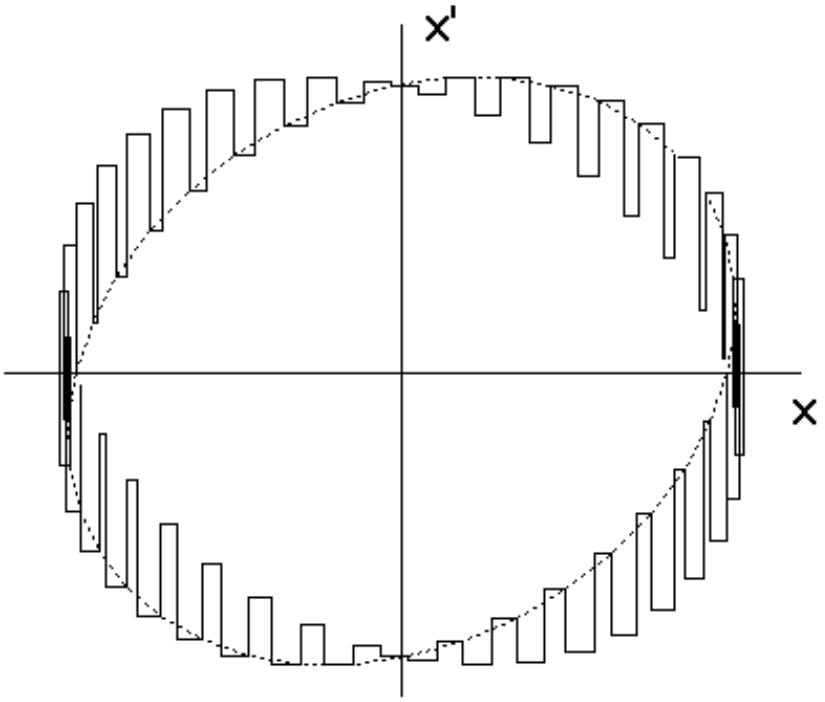


Figure 2.6: One betatron oscillation in the transverse phaseplane

focalisation cells, the displacement rises or falls linearly with time. Again the product is a straight lined path about the origin.

The effect of the **bending** is not meant to have influence on any these phaseplanes, but due to imperfections of the uniform magnetic fields, this is inevitable. Of course, if bending magnets with gradients in the magnetic field are used then, as described in the weak-force focusing system, it would have an effect on the transversal planes as well.

The **non-active straight sections** do not change the velocity or energy. In all planes, this can be described by a straight horizontal line, hence no changes in energy nor angle.

The amplitude of the motion in each of the 3 phaseplanes can be described in terms of the emittance, ϵ. This is the area that the particles surrounds, by their path (measured in $[\pi\, mm\, mrad]$). It is similar to the standard deviation, σ, of the particle from the closed orbit. All particles described in the same phaseplane, gives the total emittance of the beam. The emittance is inverse proportional to the momentum, so when accelerated, the emittance decreases as $\frac{1}{p}$. This is called **adiabatic damping**. However,

as a consequence of decreasing the momentum, as in the AD, the emittance rises in stead. The normalised emittance is given as $\epsilon_N = \beta\lambda\epsilon$, and is independent of the momentum. According to the theorem of Louville, this normalised emittance is conserved as long as no cooling is performed. The 95% emittance of a beam, is defined to contain 95% of the particles. That is, 95% of the paths can be drawn within this area, see figure 2.7. Typical values for the emittance of the beam, in AD lattice, is in the neighbourhood of 5 $[\pi\, mm\, mrad]$.

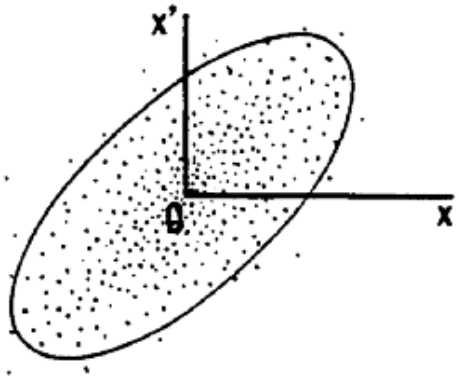


Figure 2.7: Snapshot of Particles in the Horizontal Transverse Phaseplane

2.4.2 Beam instabilities

The beam instabilities is a science of its own. Quite complicated matters can make the beam instable and it would be outside the scope of this project, to cover them profoundly. A short introduction to the most important causes of instabilities is introduced here.

In the preceding section, the paths were described as were they only determined by the effects of magnetic fields. There is another effect from the residual gases in the imperfect vacuum of the beam chamber. There are still collisions with these gas molecules, which results in scattering of the particles. Three types of collisions can occur , single coulomb, multiple coulomb and nuclear scattering. They all contribute to disjunctive changes of the movement about the phaseplanes. If their changes are too big, then the system is not able to recover the control of the particles. Smaller variations are not fatal, due to the beam control actions mentioned above.

A far more important factor is beam resonance, this leads to loss of the whole of the beam. The resonances occur when the beam fulfils the equation

$$nQ = p$$

- where n and p are integers. The lowest order resonances, low n values, are the strongest and most destructive. As the order rises the importance declines and higher order than 5 can, as a rule of thumb, be neglected.

The combination of horizontal and vertical tunes can lead to resonance, as well. This occurs in the same way, when the tunes reach values which fulfils the following equation.

$$nQ_h + mQ_v = p$$

- where n, m and p are integer values.

In a plane with the horizontal and vertical tunes as axis', the instability lines takes shape as on figure 2.8. The single resonant is shown up to 5th order, with solid lines. The combined up to 3rd order with dotted. The position of the point (q_h, q_v) should be far from resonances as possible, thus placed far from lines. The Schottky analysis allow us to zoom in on a square which is 0.1125×0.1125, shown as punctured line on figure 2.8. The q values are presumed to be in the middle of this square. They should not coincide with the strong 3rd order resonance, at 5.33, nor the 5th order, at 5.4. These are the most important ones to avoid.

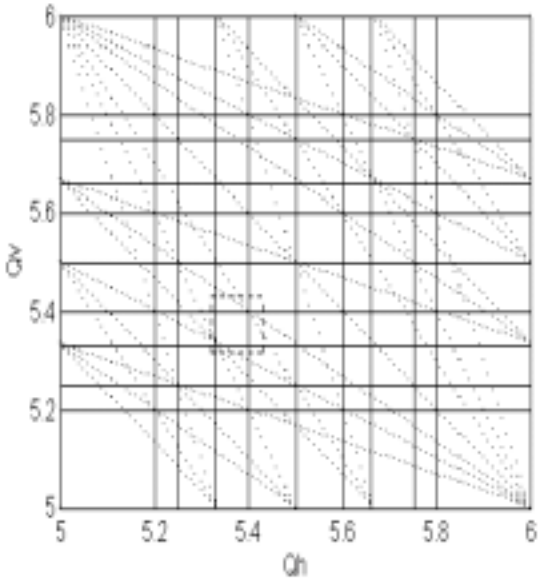


Figure 2.8: Resonant tunes

The reason why this equation leads to instability is, that the beam has a tune that becomes a fraction of the revolution frequency. This means subsequent passages of the same spots. A tiny imperfection is then accumulated thousands of times and even a tiny contribution, will in an instant, result in an unstable non-recoverable oscillation.

The Q values can be changed by adjusting the quadrupole magnetic field. The Schottky analysis is an important tool for this adjustment. It is important to know the current q value, in order to know what to change. To begin with, the q values are given by calculations and complex simulations. The values are currently at, $Q_h = 5.39$ and $Q_v = 5.37$.

2.4.3 Matrix representation

A frequently used method of calculations on a synchrotron lattice optics is to divide each section or cell independently. The parameters are the incoming displacements and velocities, say $[x_0, \frac{dx_0}{dt}]$ and change of the parameters after the passage of the section or cell $[x, \frac{dx}{dt}]$. This can be represented in a matrix form as: $[x, \frac{dx}{dt}] = \overline{T}'[x_0, \frac{dx_0}{dt}]$, where $\overline{T}$ is a 4×4 transport matrix and ' is the Matlab notation for transponed. $\overline{T}$ can be a transformation matrix, consisting of either functions or simply numbers. It depends on both the architecture of the cell and the type. A single straight section with no magnetic fields has a transformation matrix as $\overline{T} = [\ 1\ l\ ;\ 0\ 1\ ]$ where again ; is Matlab notation for new row and l is the length of the section. By multiplying together all matrices in the ring, a total transformation matrix, say $\overline{S}$, for the system is calculated. This matrix should not divert when lifted to a high order, ($\overline{S}^n$ for $n \to \infty$). This would mean that the system was unstable.

Some more sophisticated methods are using vectors containing the normalised change in momentum as well, for the horizontal phase plane this becomes $[x; \frac{dx}{dt}; \frac{\Delta p}{p}]$. Such analysis' are always done by optical simulation programs.

2.4.4 AD cycle

The anti-protons, in the AD, are produced from a 26 $[GeV/c]$ beam of 10^{13} protons. The protons hit a target of iridium and after the coalition, antiprotons, at different energies, are produced. The antiprotons, with a momentum of around 3.5 [GeV/c], are collected. This is an injection of about 5 10^7 antiprotons . Their energy corresponds to a start revolution frequency of 1.587 [MHz]. This corresponds to a velocity close to the speed of light. Then the beam is bunched to avoid momentum dispersion, this dispersion is decreased from $\pm 3\%$ to $\pm 1.5\%$. The beam is then stochastically cooled to an emittance of 5 $[\pi\, mm\, mrad]$. The beam is decelerated to 2 [GeV/c], where it is cooled to avoid adiabatic beam blow up. Then again decelerated and cooled with electron cooling. The beam is then extracted from the synchrotron to experiments. The number extracted is about 1.2 10^7, thus an efficiency of about 25%, in respect to what is injected.

Combining the formulas $p = \gamma m_0 v$ and $f_r = \frac{v}{L}$ reveals

$$f_r = p\frac{1}{L}\sqrt{\frac{c^2p^2}{c^2m_o^2 + p^2}}$$

from which the revolution frequencies are calculated.

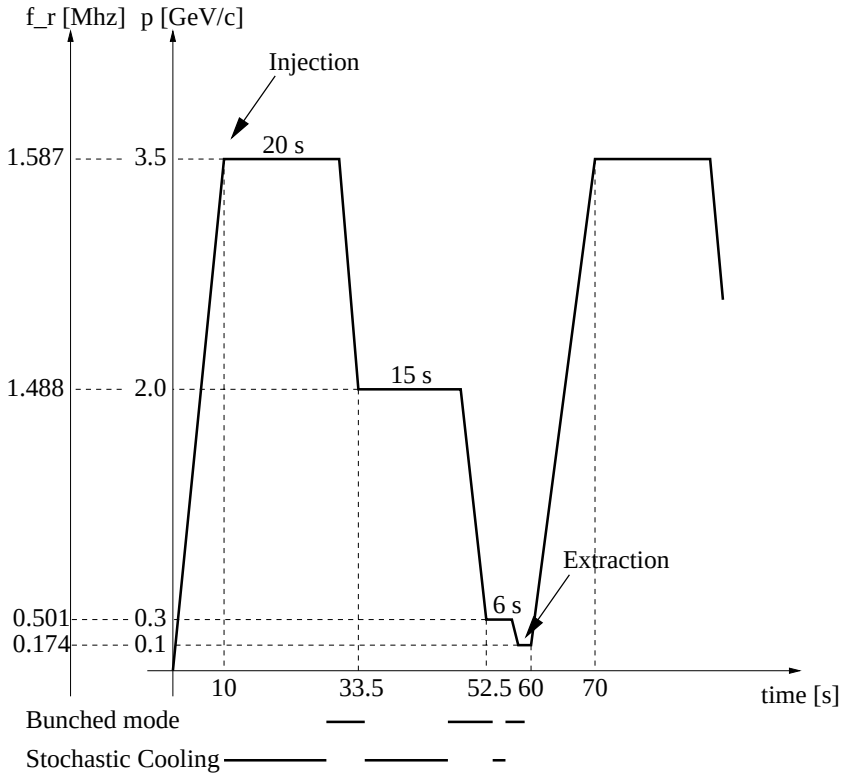


Figure 2.9: Beam state steps in the AD

Chapter 3

Schottky Noise

In this chapter, the Schottky noise will be introduced. The chapter is strictly theoretical, containing quite a lot of mathematics and signal analysis. Only the principles of signal behaviour is introduced. Considering the complete signal behaviour would involve too many parameters, to be performed in a nice analytical way. Such complete analysis' are left for the simulators to do.

At first the signal source is introduced, then the way it is detected. Following is a the main part of this chapter, going through the analysis of these four types of signal. The system noise contributions to these signals is only just mentioned. Then the signals are analysed from a power spectral density point of view. This including the effect of a noise floor. This part is statistical and considers the effect of averaging spectra. The windowing function is introduced and the effect of it, applied to the power spectral density calculation. Finally the timing of such signal detection, is gone through. This finishes with a sheet containing a draft of the analysis timing.

3.1 What is Schottky noise

The name Schottky noise signal, is a bit misleading. It is not really noise, as we are used to think of it. Ordinary noise has uncorrelated nature, whereas Schottky noise is a bit different. Schottky noise, is an addition of many coherent signals, but individually uncorrelated in phase and frequency. In our system the coherent signals appear, when the same particle passes the same pick-up successively in a systematic way. However about 50 million other particles are doing likewise, but with no correlation to each other. Special techniques is thus needed to observe the signal, in order to derive the Schottky noise information.

We do not detect single particle behaviour, but a kind of very detailed behaviour of the beam. Each passage of particles does add information that

could be analysed, if we had equipment that was fast enough. However, the particles are passing at a speed close to the speed of light, so the detected signal is an integration of a large number of particle passages. This is equal to a smearing in the frequency domain. So the individual particle components becomes a Schottky band.

Each band has a bandwidth and it rises with the harmonic number. We presume a uniform distribution of frequencies, around the revolution frequency, within Δf and we have N particles. The DC current is $N < f_{rev} >$, where $< f_{rev} >$ is the mean revolution frequency. The first harmonic broadens Δf, the next $2\Delta f$ and so on. The band around the n'th harmonic becomes $n\Delta f$. Eventually the band will overlap, as this is repeated with identical frequency harmonics and rising bandwidths. Bands with no overlap has an integral that gives information about the number of particles. This is denoted the **intensity**. The band also gives information about the spread in revolution frequencies and the geometric properties of the beam.

3.2 Signal detection

Application specific detectors, denoted **pick-ups**, are developed to detect the displacement of the beam, in respect to the synchronous particle.

The longitudinal pick-ups use the fact that a motion of charges induces a rotating current, in a surrounding material, normal to the particle motion ($\overline{J} = nq\overline{v}_p$). A beam travelling inside a tube, would result in a current at the inner surface of the tube, proportional to the number of charges in the beam, but in the opposite direction[1]. By introducing a discontinuity in the tube, the beam gets slightly disturbed, but not significantly. The current is then not able to pass the discontinuity, but is lead through an impedance in stead (a coupler). This enables measurement of the beam current. The principle of the pick-ups used in the AD project, is of this resistive-gap-type, see figure 3.1.

The transverse pick-ups rely on the principle, that a charged particle, in the middle of a large two plate capacitor, will result in the same potential on both plates of the capacitor, whereas a small displacement, will have an impact on both plates (see the later section 3.2.1). So both sign and value of the displacement, is detected this way. The detection is split up in the transverse vertical and the transverse horizontal plane.

[1]The charges are travelling in the same direction, but are of different polarity

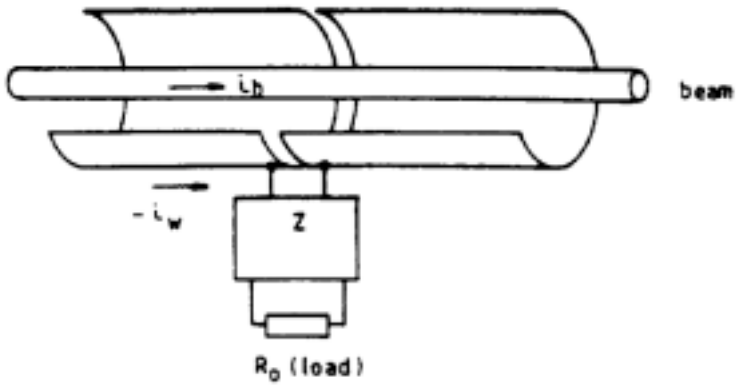


Figure 3.1: Principle of longitudinal resistive gap type pick-up

3.2.1 Charge passage of transverse pick-up

The transverse pick-up consist of two parallel plates, with bended corners. The length of the pick-up is ∼1 meter.

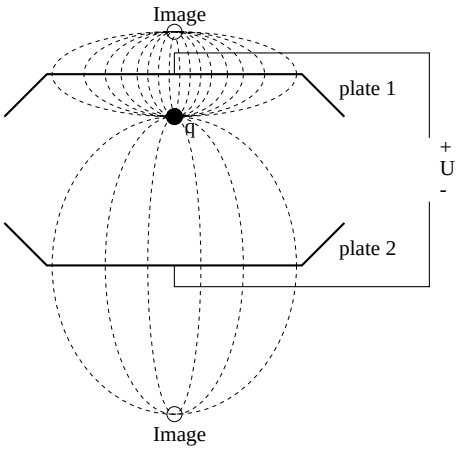


Figure 3.2: Cross section of transverse pick-up

When a charge is between the plates, it induces a charge each of the plates. This can be modelled as having two image charges at each side. The field lines crosses the plate with a 90 degree angle and creates an induced charge on the surface. As the two image charges are not of equal value, due to difference in distance from the charge, there will not be an equivalent amount of charge on both plates. The voltage between the plates will be

$$U = \frac{Q_{plate\ 1} - Q_{plate\ 2}}{C}$$

The actual induced current, is quite a difficult piece of calculation to

perform. Normally this is done numerically. A good approximation is to assume that the voltage varies linear with the displacement from the centre. If the charge is situated at the plates, then the image charge coincides with the charge and the other plate has no induced current. In the middle the induced current is zero, as both image charges are equal. In between we thus assume this linear variation leaving us with the transfer function visualised in figure 3.3. The solid circles, on the figure, shows the values where the voltage are exact.

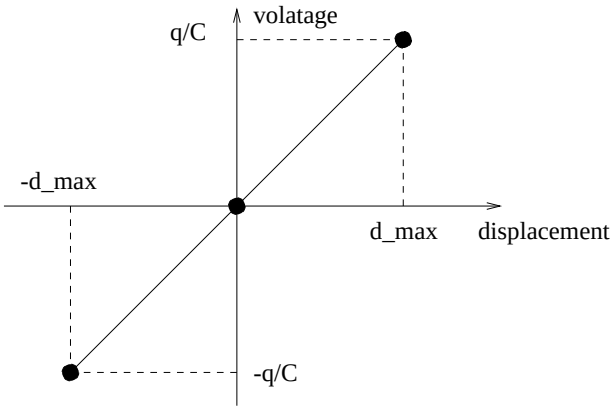


Figure 3.3: Charge displacement versus Voltage

When a charge enters the gab of a pick-up, it induces a current on both plates. When it leaves the pick-up gab, it induces a current of exact opposite polarity. The spectrum of such subsequent passage is going to be studied in the following. First we split to signal up into two signals, corresponding to the entering and leaving, of the pick-up.

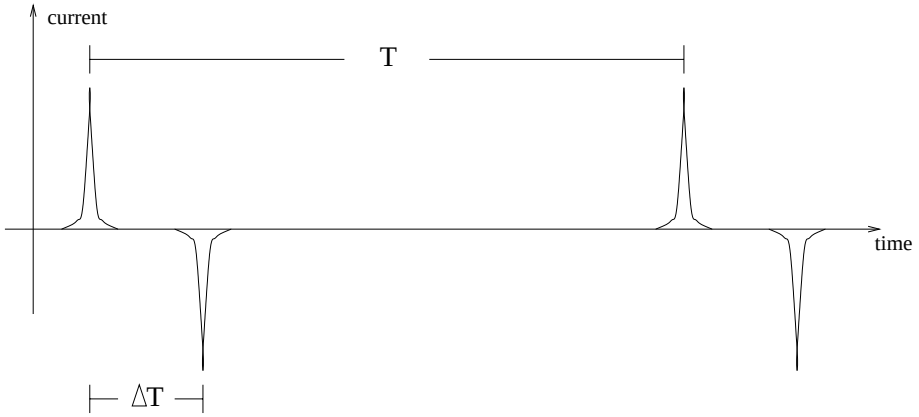


Figure 3.4: Charge passage

$$signal = a(t) + b(t)$$

The first signal, $a(t)$, is just a repeated δ-function which has a periodic spectrum of a sinc[2] function.

So $a(T)$ has the spectrum $A(f)$, written.

$$a(t) \leftrightarrow A(f)$$

The $\leftrightarrow$ symbolises the reversible Fourier transformation. The left side is the time domain and the right side frequency domain.

Then for the spectrum of $b(t)$ we have the same, only negative, and the phase of them is a bit displaced in time.

$$b(t) = -a(t + \Delta T) \leftrightarrow -A(f)e^{j2\pi\Delta Tf} = A(f)e^{j(2\pi\Delta Tf+\pi)} = B(f)$$

The signal sum can then be expressed from only the spectrum of $a(t)$ as

$$signal = a(t) + b(t) \leftrightarrow A(f)(1 + e^{j(2\pi\Delta Tf+\pi)})$$

This envelope has an absolute value that vary with the frequency from 0 to 2 and in phase from $-\pi/2$ to $\pi/2$. In a complex plane, the envelope follows the trajectory drawn at figure 3.5

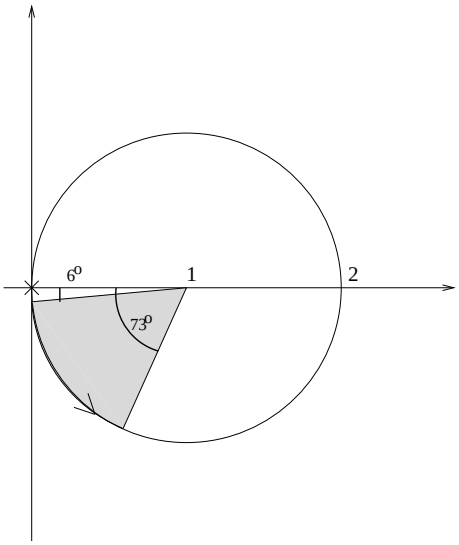


Figure 3.5: Complex envelope

[2]a function with nature, $sinc = \frac{sin(at)}{sin(t)}$

In our setup we have a 1 meter pick-up, placed in a 182 meter long lattice. This gives a constant ratio between T and ΔT of approximately 182. The time between passages, T, is inverse proportional to the revolution frequency, this frequency vary from 0.174 MHz to 1.56 MHz in the AD. A rewritten version of the envelope for our system, becomes.

$$1 + e^{j(2\pi \frac{f}{182 f_{rev}} + \pi)}$$

With these two frequency intervals, f and f_{rev}, we use only from $\sim$6 to $\sim$73 degrees, hence shaded angle interval on figure 3.5.

3.2.2 Noise

The signals are subjected to several noise contributions, before the final treatment is done. The noise contribution from the measurement system to the Schottky signals, is beyond the scope of this project to analyse. A short introduction is, however, summed up in the following.

Starting from the beam itself, there can be some misalignment in the transverse pick-up, so that the differential signal is added an offset. The power of a differential signal (Schottky signal), rises with $\sqrt{N}$ and the offset contributes with N. So misalignment, even tiny, contributes to a very strong biased signal, that exceeds the interesting Schottky signal with many factors. The AD beam consist of about 5 10^7 particles, so an offset is amplified 7000 times (77 [dB]), more than the Schottky signal. The same effect occurs when the particles are performing coherent oscillations. Such a signal will appear as a betatron signal, but with a huge amplification compared to the Schottky signal.

The head amplifier, amplifying the pick-up signal, is not ideal thus introducing errors of different kinds. First of all, there is never a complete linear amplification in all of the input interval. Some level of distortion will always be present. Secondly, the amplification depends on temperature and resistance values and this introduces a level of coloured noise.

When the signal is all set to be processed it passes an A/D converter that introduces quantisation noise at a level of $\Delta V/2^n$, where ΔV is the signal amplitude interval and n the bit resolution. Having a random signal these errors introduced are uncorrelated and random as well, thus white noise. In this case, however, the revolution frequency is close to fixed and some patterns are bound to be stable. Thereby some correlation between quantisation errors arise and larger parasitic frequency components, called spurious components, occurs.

Taking only the noise from the system amplification into account, one gets an estimate of the signal to noise ratio, available for treatment. From

the longitudinal pick-up a spectral density noise level of $1.5-2$ $[pA/\sqrt{Hz}]$ is aimed. For the transverse case the spectral density is $0.3-0.5$ $[pA/sqrtHz]$.

As the AD cycle changes state, the amplification and noise levels changes, as well.

3.3 Beam parameters

In this section the signal nature will be introduced. The detected signals are analysed in order to derive the beam parameters from them. This analysis is done with signal analysis techniques. It is mostly theoretical calculations done in the continuous time and frequency domain.

3.3.1 Signal treatment

There are four types of beam signals, that has to be considered. First of all, the beam can be either in a bunched or unbunched state. Second, there is both longitudinal and transverse signals detected in both states.

The transverse are split up in vertical and horizontal, but their treatments are similar.

The description of the signal treatment and what we can expect from it is divided into the four types described above. For additional descriptions please refer to [1] and [2]. Most of the descriptions are supported by Matlab calculations and plots.

Unbunched r.m.s. current

Most of the processing is done from unbunched signals and the calculations is almost identical for longitudinal and transverse case. The only difference is the amplitude of the detected signal.

The particle pick-up passage is estimated to be very fast compared to other time constants in the system, so a passage is modelled as a delta function, $\delta(t)$. Each passage occurs with the revolution frequency of the particle in mention, f_i, thus a train of delta pulses.

$$\begin{aligned} i(t) &= ef_i \sum_{n=-\infty}^{\infty} exp(jn\omega_i t + \phi_i) \\ &= ef_i + 2ef_i Re[\sum_{n=1}^{\infty} exp(jn\omega_i t + \phi_i)] \\ &= ef_i + 2ef_i \sum_{n=1} cos(jn\omega_i t + \phi_i) \end{aligned}$$

In the frequency domain, a train of delta pulses has an amplitude spectrum with value ef_i at DC and each at harmonic[3] nf_i.

Detection of several particles reveals just an addition of everyone of them. They all have random phases in relation to each other, which means that the addition is only non-zero for the DC value. Deriving the r.m.s value of the first harmonic band reveals a different spectrum.

$$\begin{aligned} i_{rms}(f_{band}) &= \sqrt{< (2e(f_{rev}+\Delta f_1)cos(2\pi f_1 t+\phi_1)+...)^2 >} \\ &= 2e\sqrt{< ((f_{rev}^2+2f_{rev}\Delta f_1+\Delta f_1^2)cos^2(2\pi f_1 t+\phi_1)+... \\ &\quad 2(f_{rev}^2+\Delta f_1\Delta f_2+f_{rev}(\Delta f_1+\Delta f_2))\times \\ &\quad cos(2\pi f_1 t+\phi_1)cos(2\pi f_2 t+\phi_2)+...) >} \\ &\approx 2ef_{rev}\sqrt{< cos^2(2\pi f_1 t+\phi_1)+cos^2(2\pi f_2 t+\phi_2)+... >} \\ &= 2ef_{rev}\sqrt{\frac{N}{2}} \end{aligned}$$

In the first equation above a frequency, of the revolution frequency plus a correction, is introduced. The equation is squared, thus resulting in clean particle squares and products of different particles. The products from different particles cancel out, due to the random phase factors. The clean square is multiplied with the amplitude $(f_{rev}^2+2f_{rev}\Delta f_1+\Delta f_1^2)$, which is reduced to just f_{rev}. The second term is zero in mean and the third so small compared to the revolution frequency, that it is neglected. This leaves us with the third equation. The mean value of a squared cosine is just $1/2$, so having N of those we get the last equation.

The r.m.s. value of the current, is not measured from time signals, as in the formulas above. It is measured via their Fourier transform. The square root of an integral of a PSD function is equal to the r.m.s. current of the integral interval. This PSD function is found from squaring the absolute value of Fourier transformation and dividing it by $T/2$. The power spectral density estimate in a frequency interval, from m_{min} to m_{max} is thus.

$$\hat{G}_x = \frac{2}{T}\sum_{m=m_{min}}^{m_{max}} |X(m)|^2$$

[3]Some physicists prefer a spectrum with only positive frequencies why the amplitude of the harmonic becomes $2nf_i$

3.3.2 Unbunched beam longitudinal decomposition

Momentum spread

Due to spread in the momentum, Δp, there is a spread in the synchrotron frequencies, Δf. From measuring the Schottky bandwidth of a harmonic band, hence $< \Delta f >_{rms}$, the momentum spread can be calculated.

Intensity

The total amount of particles is proportional to the squared r.m.s. current per band, $i_{rms}^2 \propto N$. Every harmonic Schottky band has the r.m.s. current, $i_{rms} = 2ef_{rev}\sqrt{\frac{N}{2}}$, from which the amount of particles, the intensity, can be calculated.

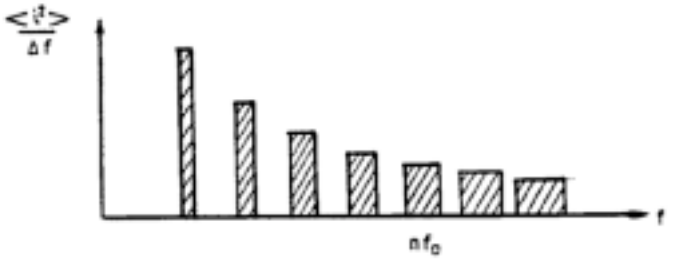


Figure 3.6: Unbunched Longitudinal Frequency Spectrum

3.3.3 Bunched beam longitudinal decomposition

From this signal we only measure the intensity, by measuring the values of two harmonics according to the principle explained below. This section is also introduces the effect of synchrotron oscillations on the spectrum even though it isn't used for parameter estimation.

Measuring intensity

The spectral components at harmonic frequencies, are in theory the same for any harmonic, as mentioned before. However, this is not entirely true in the real world. In section 3.2.1 the theory is derived for the transverse pick-up, but the principle applies to the longitudinal as well. Say that only two particles are detected and they have the same revolution frequency. This is almost true for every particle, only they have different phases.

$$a(t) + a(t + \Delta T) \leftrightarrow A(f)(1 - e^{j(2\pi\Delta T f)})$$

This is the effect that damps higher harmonics. We want to know only the DC value from the longitudinal unbunched signal, but this band is not fitted for measuring. In stead we measure at a frequency f and $2f$. We don't know the ΔT, but from measuring two harmonic values we don't need to. Say we have two measurements, X_1 and X_2 and we want to find the value A(f).

$$X_1 = A(f)(1 - e^{j(2\pi\Delta T f)}) = A(1 - K)$$
$$X_2 = A(f)(1 - e^{j(2\pi\Delta T 2f)}) = A(1 - K^2)$$

Having these two equations with two parameters, we can eliminate K and get an equation for the A, which is half the DC-value.

$$A = \frac{X_1^2}{2X_1 + X_2}$$

Solving this equation will thus reveal the estimated DC value, without measuring anything even close to this noisy band.

Fourier transforming the synchrotron oscillation

When the beam is bunched then the revolution frequency of each particle is f_{rev}, in mean, but with an oscillation about this frequency. This is similar to an equidistant distribution of pulses, but slightly sinusoidal displaced in time. The same problem is known from signal processing as an inaccuracy/error in the moment of sampling. This is similar to a signal of the nature, $cos(anT + bsin(cnT + d))$, which has the frequency spectrum shown on figure 3.7

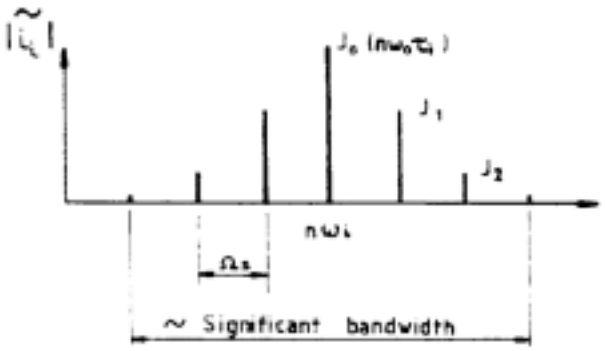


Figure 3.7: Complex of Synchrotron Satellites

The signal can be modulated as series of odd distributed delta functions.

$$i(t) = ef_0 \sum_{n=-\infty}^{\infty} \delta(t - nT - \tau_i sin(\Omega_s t + \Psi_i))$$

by Fourier analysis this can be transformed into:

$$i(t) = ef_0 + ef_0 Re[\sum_{n=-\infty}^{\infty} exp(jn\omega_0(nT + \tau_i sin(\Omega_s t + \Psi_i))]$$

Using the relation $exp(jzsin\theta) = \sum_{p=-\infty}^{\infty} J_p(z)exp(jp\theta)$, where J_p is the Bessel function of first kind, the following equation is obtained.

$$i(t) = ef_0 + ef_0 Re[\sum_{n=-\infty}^{\infty} (e^{-j2\pi n\frac{f}{f_{rev}}} \sum_{p=-\infty}^{\infty} J_{|p|}(n\omega_0\tau_i)exp(j(n\omega_0 t + p\Omega_s t + p\Psi_i)))]$$

Taking only one single harmonic of this current:

$$i_n(t) = 2ef_0 e^{-j2\pi n\frac{f}{f_{rev}}} Re[\sum_{p=-\infty}^{\infty} J_{|p|}(n\omega_0\tau_i)exp(j(n\omega_0 t + p\Omega_s t + p\Psi_i))]$$

We see that this is just a frequency component, at $n\omega_0 t + p\Omega_s$, with an amplitude of $ef_0 e^{-j2\pi n\frac{f}{f_{rev}}} J_{|p|}(n\omega_0\tau_i)$.

Every harmonic is a complex of several distributed frequency components, denoted satellites. Their amplitudes decreases significantly, with distance from the particles mean revolution frequency, $p = 0$, especially for low harmonics. Already the first or second harmonic is almost zero, in the example showed below. The tail(hence frequency components for $p = 1, 2, 3.$) gets longer with the harmonic number, n, whereas the centre component gets smaller in a damped oscillating way (see fraction of Bessel function on figure 3.8). One would think that, as the tails grew, the harmonics would eventually overlap each other. But the distances between successive satellites is the synchrotron frequency, Ω_s, and the distance between complexes of satellites is the revolution frequency, ω_0. If the revolution frequency is much larger than the synchrotron frequency, $\omega_o >> \Omega_s$, then this overlap will be insignificant.

In a Matlab simulation the satellite complexes is obtained by creating signals with a varying longitudinal displacement. It is seen from the plot, figure 3.9, that the component at 1.65 $[MHz]$ is a bit larger than the one at 13.5 $[MHz]$, this is due to the overlap, from higher harmonics, mentioned above. The synchrotron frequency used for this simulation is 0.15 $[MHz]$ and the revolution frequency is 1.5 $[MHz]$. Apart from this overlap, the value corresponds to what must be expected, hence table 3.1.

In practical these Schottky spectra can not be measured. The signal from a bunched longitudinal beam is too strong and the amplifiers can not detect the fine differences between particles. In stead the amplitude values of two harmonics are measured, one at, say m, and the other at 2m. By this the DC value, of the beam, can be calculated. This is possible by knowing the envelope function, which are derived in section 3.2.1.

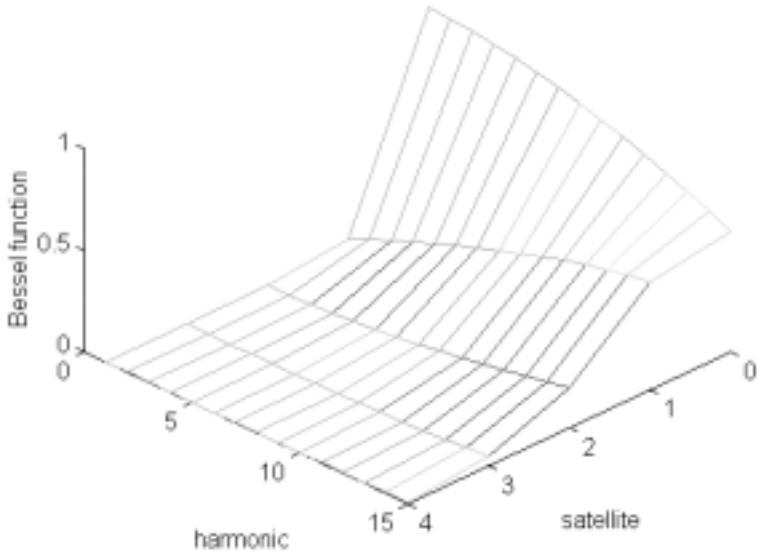


Figure 3.8: Bessel function, $J_p(n\omega_i\tau_i)$

Table 3.1: Satellite amplitudes

Centre frequency	$J_0(2\pi\ 0.1) < signal >$	0.9037*0.0269	0.0243
First satellite	$J_1(2\pi\ 0.1) < signal >$	0.2989*0.0269	0.0080
Second satellite	$J_2(2\pi\ 0.1) < signal >$	0.0477*0.0269	0.0013
Third satellite	$J_3(2\pi\ 0.1) < signal >$	0.0050*0.0269	0.0001

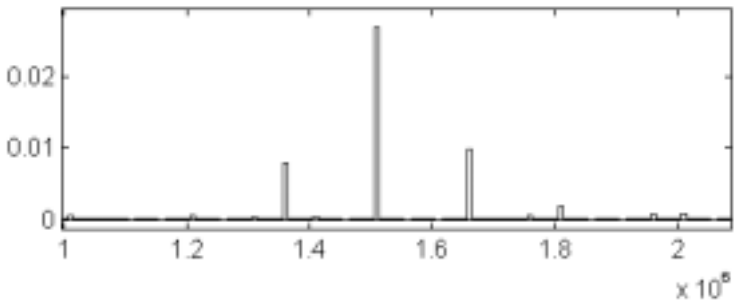


Figure 3.9: Satellite generated from signal simulation in Matlab

The DC value is proportional to the number of particles in the beam. In this mode the beam is decelerated, so the number of particles is surveyed in case of loss.

3.3.4 Unbunched beam, transverse decomposition

Single particle

In the transverse plane, the particles are performing betatron oscillations with a frequency of Qf_i. The signals from a single particle detected by a pick-up, can be simulated as, the betatron sine wave sampled with the revolution frequency of the particle.

$$i_i(t) = sinewave * revolution = a_i cos(2\pi f_i q_i t) * \delta(t - nT_i)$$

Where a_i is the amplitude of oscillations, T_i the time of revolution and q_i the fractional part of Q. The fractional part is used due to the fact that the integer number of oscillations, in between discrete pick-ups, can not be detected. For a single particle this is similar to a sine wave, with a betatron frequency sampled with the revolution frequency. Thus a frequency spectrum with lines of half the sine amplitude at harmonics, $f_i q_i$, $f_{rev} - f_i q_i$, $f_{rev} + f_i q_i$,...Again simulated in the Matlab model, this reveals a frequency spectrum as seen on figure 3.10.

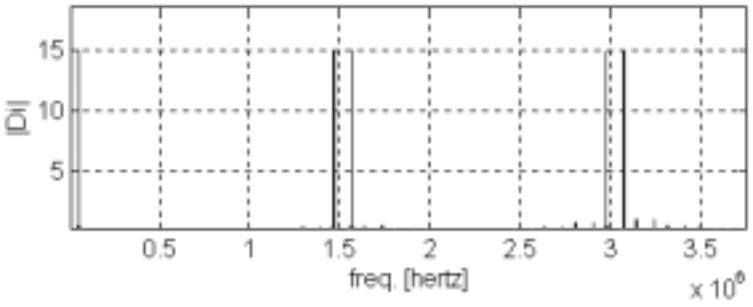


Figure 3.10: Transverse Frequency Spectrum of Single Particle

In this case simulated with a revolution frequency of $1.51[MHz]$ and a betatron frequency of $50[kHz]$.

For N particles an analogy can be drawn to the calculations done on the unbunched beam, in the longitudinal case. The theory was derived in section 3.3.1 and the only difference is an introduction of an amplitude. The amplitude is introduced as $a_i = a_{rms} + \Delta a_i$ and reduced as done with the revolution frequency.

Here, as well, we get a spread in the frequencies and each and every band containing an r.m.s. current of, $i_{rms}(f_{band}) = a_{rms} e f_{rev} \sqrt{\frac{N}{2}}$.

Several particles

When extending to many particles, we must take into account that the particles have slightly different revolution and betatron frequencies. The

revolution frequencies of the particles are varying about the mean revolution frequency, between $f_0 \pm \frac{1}{2}\Delta f$. The q values vary Δq_i about some mean q and the resulting betatron frequency, for a given particle, is $q_i f_i$.

To illustrate how the spectrum looks like, when both f_i and q_i varies, we let f_i be constant and calculates the spectrum for a given interval of q. Then repeating this, for several revolution frequencies, we can add our spectra together and get an idea of how it looks like, for varying f_i.

In the sketch, shown on figure 3.11, this is done with three constant revolution frequencies. It is seen that in the cases of constant revolution, we get spectra that looks as one would expect them to, hence single particle spectrum shifted at each side of the revolution frequency. But the two total harmonic spectra, consisting of the addition, has different Schottky bandwidths. By comparing such adjacent Schottky bands, it is possible to calculate Δq_i of a beam.

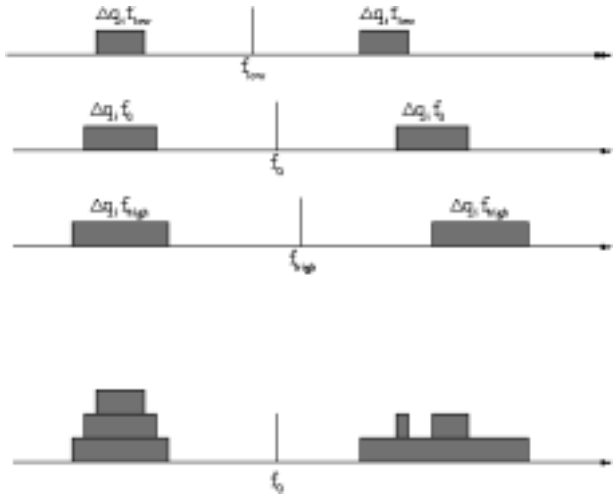


Figure 3.11: Example of different bandwidth of bands

The bandwidth of two adjacent Schottky bands are $\Delta f = (n \pm q)\Delta f_i \pm 2\pi f_0 \Delta q_i$.

If it is possible to identify some identical patterns in both Schottky bands, say a large peak or complex, then the q value can be determined from either the peak or the complex. The q value for a peak would be interpreted as being several particles having this (or close to this) betatron oscillation frequency.

The estimated q value, for the betatron frequency, is the centre of a complex of particle frequencies. This is the quantity calculated from the unbunched transverse PSD spectra. The momentum spread is found from other measurements, thus the frequency spread can be calculated and the q value calculations can be corrected.

Emittance

Thus when each and every band contains the r.m.s. current,

$$i_{rms}(f_{band}) = a_{rms} e f_{rev} \sqrt{\frac{N}{2}}$$

, knowing f_{rev} and N enables calculation of a_{rms}. The signal from the transverse pick-up is aimed to have a zero offset, so the mean value of a beam passage, would be zero, $< i(t) >= 0$. For the emittance, ϵ, it is given that the maximum amplitude, for a single particle, has the relation:

$$\epsilon_{2\sigma} = \frac{4\sigma^2}{\beta}$$

So knowing the trajectory in the phaseplane for every particle enables calculation of the total emittance. The maximum amplitude is not found from the r.m.s. spectra, only the r.m.s. value of the amplitude. The relation between the variance, of the amplitudes, $\sigma^2\{a(n)\}$, and the r.m.s. amplitude, a_{rms}, allowing negative $a(n)'s$, is:

$$\begin{aligned} \sigma^2\{a(t)\} &= \frac{1}{N}\sum_{n=1}^{\infty}(a(n) - E\{a(n)\})^2 \\ &= \frac{1}{N}\sum_{n=1}^{\infty} a^2(n) \\ &= rms^2\{a(n)\} \\ &= a_{rms} \end{aligned}$$

So with a normal distribution of $a(n)'s$, 2σ contains 95.5% of the amplitudes.

$$\begin{aligned} \epsilon_{2\sigma} &= 4\frac{a_{rms}}{\beta} \\ &= 4\sqrt{2}\frac{L}{ce}\frac{i_{rms}}{\sqrt{N}} \end{aligned}$$

3.3.5 Bunched beam, transverse decomposition

The transverse pick-up senses both the transverse betatron oscillations and the longitudinal displacement due to synchrotron oscillations. Thus there is both the effect of oscillating signal amplitudes and an oscillating time interval of passages. These two effects are combined in the equation.

$$d_i(t) = a_i cos(q_i\omega_0 t + \phi_i) e f_0 Re[\sum_{n=-\infty}^{\infty} exp(jn\omega_0(t + \tau_i sin(\Omega_s t + \Psi_i))]$$

The last part is the same as for the longitudinal case only difference here is a cosine wave multiplied by the expression already calculated in a former section. In a frequency domain this is just a convolution between a discrete frequency component of the cosine, at $q_i\omega_0$, of amplitude $\frac{q_i}{2}$ and the former obtained frequency spectrum. Again in a Matlab simulation, this takes the form seen on figure 3.12:

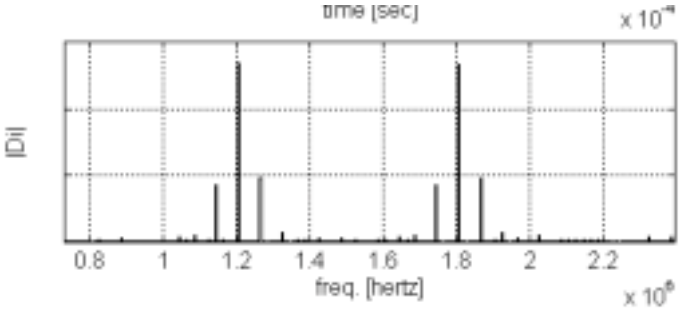


Figure 3.12: Frequency spectrum of a bunched beam in the transverse plane

In this mode we calculate the same parameters as before, q and ϵ, when the beam was unbunched. But as the particles are passing the pick-up, at the same time, a common displacement from the closed orbit will result in a great bias of the signal. This bias might saturate the amplifier so that our signal information is lost. In such a case the system can measure the parameters in a BTF[4] mode in stead. This mode is described in section 3.7.

3.3.6 Parameter calculation summary

The seven parameters calculated by signal processing is summed up on table 3.2, where $f_{50\%}$ is the centre frequency of a Schottky band and n the harmonic number.

3.4 PSD estimation

When we look at the power spectral density(PSD), of our signal, then it looks something like the sketch at figure 3.13. Actually, there is a small curvature of the noise floor, which is better estimated as a line with a derivative. The calculations done in this section however assumes a straight horizontal line, but this does not significantly effect the results obtained.

[4]Beam Transfer Function

Quantity	Calc. from	Relation	Beam State
N	i_{rms}	$N = \frac{2i_{rms}^2}{e^2 f_{rev}^2}$	Un./ Long.
N	$DC\,amplitude$	$N = \frac{2\|DC\|^2}{e^2 f_{rev}^2}$	Bu./ Long.
$< \Delta p >_{rms}$	$< \Delta f_{rev} >_{rms}$	$< \Delta p >_{rms} = \frac{p < \Delta f_{rev} >_{rms}}{(n\eta < f_{rev} >)}$	Un./ Long.
q_V	$f_{50\%}$	$q = \frac{f_{50\%}}{< f_{rev} >}$	Bu.+ Un./Trans.
q_H	$f_{50\%}$	$q = \frac{f_{50\%}}{< f_{rev} >}$	Bu.+ Un./Trans.
ϵ_V	i_{rms}	$\epsilon_{2\sigma} = 4\sqrt{2}\frac{L}{ce}\frac{i_{rms}}{\sqrt{N}}$	Un./Trans.
ϵ_H	i_{rms}	$\epsilon_{2\sigma} = 4\sqrt{2}\frac{L}{ce}\frac{i_{rms}}{\sqrt{N}}$	Un./Trans.

Table 3.2: Parameters calculated

The power spectral density estimate is found from Fourier transforming the detected signal and then multiplying with a constant according to the equation:

$$\hat{G}(k) = \frac{2f_s}{N} \, |\hat{X}(k)|^2$$

And the estimated power in a band, is just an addition of the estimated power spectral density components in this very band.

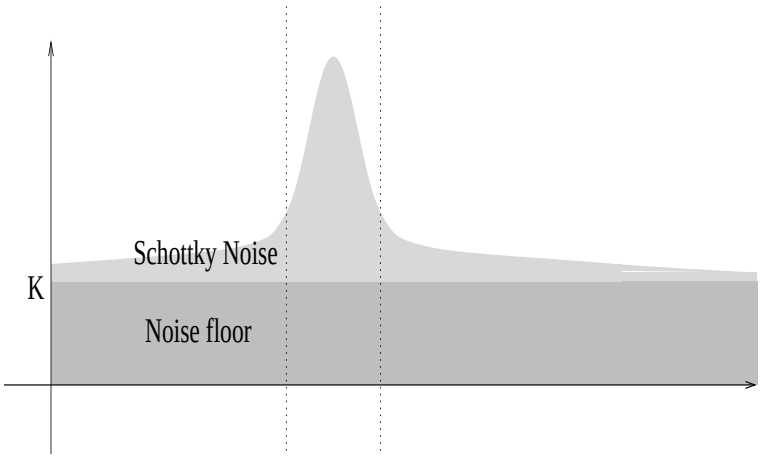


Figure 3.13: PSD of Schottky Noise Signal and Noise Floor

Schottky noise distribution

We define the signal to noise ratio to be calculated in a band, bounded by $\pm 2\sigma$. This is two standard deviations, which contains 95.45% of the Schottky signal. This is sketched on the figure as vertical punctured lines.

The signal to noise ratio is the ratio of the two areas between the vertical lines. For these calculations we freeze the Schottky distribution and vary the power of the noise floor level with the SNR.

$$SNR = \frac{0.9545}{\mu_{Noisefloor}}$$

Each PSD component of the Schottky noise, has a statistical variation about the theoretical distribution. We assume that each component is right in mean, but with a standard deviation of a certain percentage, k. Each component is then distributed as.

$$P \in N(\mu_{Schottky}, k\mu_{Schottky})$$

- and the mean values for all of these components vary with the Schottky distribution after the formula.

$$P_{Gaussian} = \frac{1}{\sigma\sqrt{2\pi}}\, e^{-\frac{1}{2}(\frac{x-\mu}{\sigma})^2}$$

This function can not be analytically integrated. Numerically it is found that the integral between $-\sigma$ and $+\sigma$ is 0.9545. The integral of the squared function in the same interval sums up to 0.2808.

We only have discrete values of this function because the frequency interval is always split up in 256 or 512, depending on the FFT length. So we have a finite array of constants which corresponds to the distribution of the mean values.

The sum of the Schottky power spectral density components with different distributions can be rewritten to be a sum of equally distributed components. The rule for a normal distributed variable multiplied with a constant is.

$$X \in N(1, \sigma) \;\Rightarrow\; cX \in N(c, c\sigma)$$

$$\begin{aligned} P_{sum} &= x_1 + x_2 + ... + x_N \\ &= c_1 x + c_2 x + ... c_N x \\ &\in N(\sum_{i=1}^{M} \mu_{Schottky,i}, \sqrt{\sum_{i=1}^{M} \sigma^2_{Schottky,i}}) \\ &\in N(\sum_{i=1}^{M} c_i, k\sqrt{\sum_{i=1}^{M} c_i^2}) \end{aligned}$$

Noise floor distribution

The noise floor is rectangular distributed in frequency, as sketched on figure 3.13. The spectral components are then in mean equal to this value and we presume that the standard deviation is proportional to this by a factor, h. The power spectral density components are distributed as:

$$P \in N(\mu_{Noisefloor}, h\mu_{Noisefloor})$$

Noise floor estimation

We would like to estimate this noise floor, to be able to subtract it from our Schottky band power estimation. We do this by taking a lot of samples, N, of only the noise floor. By doing this we can estimate the mean noise floor power, $\mu_{Noisefloor}$, with a variance of

$$\sigma^2_{Noisefloorestimate} = \frac{\sigma^2_{Noisefloor}}{N}$$

Then we can subtract the estimated noise floor power from the total measured power.

$$P_{measured} = P_{Schottkynoise} + P_{Noisefloor} - \hat{P}_{Noisefloor}$$

We allmost never get an exact estimate of the noise floor. If we want to make an analysis, of the noise floor estimate, that holds for 95.45% of the cases, then we need to consider a standard deviation of 2σ. So the noise floor contribution, when subtracted the offset, is set to be:

$$\epsilon_{Noisefloor} \leq 2\hat{\sigma}_{Noisefloor} = 2\frac{h\mu_{Noisefloor}}{\sqrt{N}} = 2\frac{h0.9545}{SNR\sqrt{N}}$$

This error is added every Schottky spectral component, so it is important to minimise this contribution. We set this to the maximum, in order to perform a worst case analysis. Then we know that 95.45% of the cases are better than estimated, thus rarely worse.

Total measured signal

The signal measured is then both Schottky noise and compensated noise floor.

$$P_{measured} \in N(M2\frac{h0.9545}{SNR\sqrt{N}} + \sum_{i=1}^{M} c_i \ , \ \sqrt{Mh^2\frac{0.9545^2}{SNR^2} + k^2\sum_{i=1}^{M} c_i^2})$$

From this it is clear that our estimate, of the Schottky band power ,is almost always biased. There are three parameters which can make this

bias less significant. First of all we need a large signal to noise ratio, which is the best way to cope with the problem. Another way is to begin the accumulation of samples as close to the Schottky band as possible, thus minimising M. The last method is to estimate the noise floor with as many samples as possible, raising N.

Estimate inaccuracy

How much is estimate biased and how significant is the standard deviation of the estimate ? The two quantities tells us how reliable the estimated power of the Schottky band is.

$$\begin{aligned}
\epsilon_{bias} &= \frac{bias}{\mu_{Scottky-power}} \\
&= \frac{M2\frac{h0.9545}{SNR\sqrt{N}}}{\sum_{i=1}^{M} c_i} \\
&= \frac{2Mh}{SNR\sqrt{N}} \\
\epsilon_{\sigma} &= \frac{\sigma_{Scottky-power}}{\mu_{Scottky-power}} \\
&= \frac{\sqrt{k^2 \sum_{i=1}^{M} c_i^2 + Mh^2 \frac{0.9545^2}{SNR^2}}}{\sum_{i=1}^{M} c_i} \\
&= \frac{\sqrt{k^2 0.2808 + \frac{Mh^2 0.9545^2}{SNR^2}}}{0.9545}
\end{aligned}$$

Examples

If we set the Schottky noise and noise floor standard deviation to be 10% of the mean value, then this is equal to $h = k = 0.1$. The sample lengths are set to respectively 256 and 512, $M = 256$ or $M = 512$, and we measure the noise floor from 4096 samples, $N = 4096$. The signal noise ratio set to either 10 or 100. This reveals.

$$\begin{aligned}
\epsilon_{bias} &\quad (SNR = 10, size(FFT) = 256) = 0.080 \\
\epsilon_{bias} &\quad (SNR = 1000, size(FFT) = 256) = 0.001 \\
\epsilon_{bias} &\quad (SNR = 10, size(FFT) = 512) = 0.160 \\
\epsilon_{bias} &\quad (SNR = 1000, size(FFT) = 512) = 0.002 \\
\epsilon_{\sigma} &\quad (SNR = 10, size(FFT) = 256) = 0.169 \\
\epsilon_{\sigma} &\quad (SNR = 1000, size(FFT) = 256) = 0.055 \\
\epsilon_{\sigma} &\quad (SNR = 10, size(FFT) = 512) = 0.233
\end{aligned}$$

$$\epsilon_\sigma \quad (SNR = 1000, size(FFT) = 512) = 0.055$$

From this is seen that if our signal to noise ratio is not significantly high, then we can't rely on the result. The bias error can be improved by taking more samples of the noise floor, whereas the standard deviation is unaffected by this.

It should be kept in mind that this analysis is theoretical and only describes the mechanisms, the numbers are thus not estimates of real errors. Estimates of real errors requires knowledge of real parameters and these are not known at the moment. They can only be found by statistical studies of the AD beam nature. The analysis made is however fully applicable, when it comes to deciding the importance of the different parameters.

3.5 Effect of Windowing

All of the above calculations are carried out on ideal signals, unaffected by the processing that the signals are submitted to. However the signals are NOT ideally transformed into their frequency spectra. First of all, the FFT algorithm presumes that the signal is periodic, with the FFT length. If this is the case, then the algorithm is precise. The assumption implicitly made by the FFT algorithm is sketched in figure 3.14. The solid line represents the real input data and the dashed line the extrapolation assumed by the algorithm.

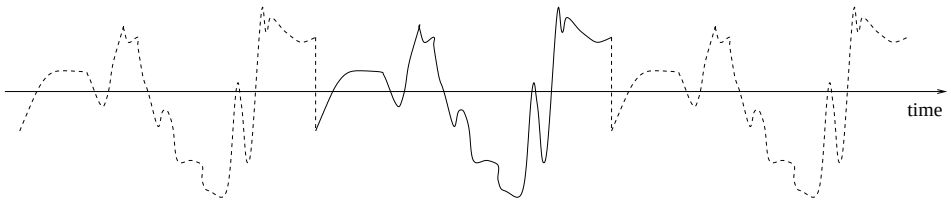


Figure 3.14: Signal assumed by the FFT algorithm

For a Schottky noise signal this assumption is wrong. There is no such periodicity. When there is not a periodicity, there will be a "discontinuity"[5] in the assumed signal input. This induces frequency components that are not present in the signal, but originates from the abrupt transition from one end to the other.

To avoid this abrupt change, we can multiply with an envelope function that declines towards zero, near the ends of the signal input vector. Such a

[5]Discontinuity is put in quotation marks because a discrete signal, by definition, always is discontinuous, but not necessarily with these large jumps of subsequent signal values

function is denoted a window function and is normally used when Fourier transforming a discrete signal.

On figure 3.15 the difference between windowed and non-windowed Fourier transformations, is shown. Note that the windowed versions get smeared a little, but not significantly, whereas the non-windowed is either precise or quite smeared.

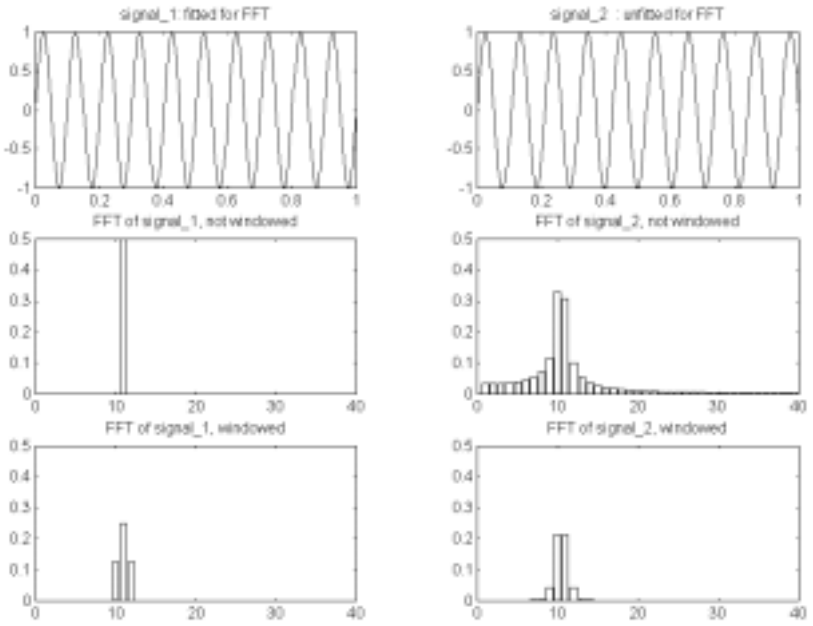


Figure 3.15: Consequence of windowing before performing FFT

Different window functions

A straight-forward window function, is the rectangular window which has the spectrum.

$$H_{square}(f) = a\frac{sin(\pi f N \Delta T)}{sin(\pi f \Delta T)}e^{-j\pi f(N-1)\Delta T}$$

This frequency spectrum has a good mainlobe, in fact the best we can get from the chosen number of coefficients. The sidelobes, however, are quite wide this is the price we pay to cut off the signal so sudden. We can however do it softer, by using other window functions such as Hamming, Hanning, Blackman, Bartlett or even a home constructed one. These have frequency spectra with less significant side lobes but wider main lobes (see figure 3.16). At figure 3.15 a Hanning function was used

The analogue spectrum of these, above mentioned window functions, can be calculated by use of the spectrum of the rectangular window function, in combination with the rule of frequency shifting.

$$e^{j\omega_0 n}x[n] \leftrightarrow X(e^{j(\omega-\omega_0)})$$

The Hamming ($a = 0.54$ $b = 0.46$) and Hanning ($a = 0.50$ $b = 0.50$) window are easily calculated as.

$$\begin{aligned} H(f) \quad &= ae^{-j\frac{\omega M}{2}}\frac{sin(\frac{\omega(M+1)}{2})}{sin(\frac{\omega}{2})} \quad - \frac{b}{2}e^{-j\frac{(\omega-\frac{2\pi}{M})M}{2}}\frac{sin(\frac{(\omega-\frac{2\pi}{M})(M+1)}{2})}{sin(\frac{(\omega-\frac{2\pi}{M})}{2})} \\ &\quad - \frac{b}{2}e^{-j\frac{(\omega+\frac{2\pi}{M})M}{2}}\frac{sin(\frac{(\omega+\frac{2\pi}{M})(M+1)}{2})}{sin(\frac{(\omega+\frac{2\pi}{M})}{2})} \end{aligned}$$

The rectangular Hamming and Hanning window spectra, are drawn at figure 3.16. At the dB scale, looking at the side lobes, the rectangular window is the largest, then the Hamming and at last the Hanning. The window chosen for this application is Hanning because the side lobes, close to the mainlobe, is damped the most.

The effects of windowing is smearing of the true signal spectrum because of the convolution with the window spectrum. This has two effects, first two distinct frequencies can merge, if they are closely placed to one another (loss of resolution) and second a single frequency is affected by sidelobes from the surrounding frequencies (leakage).

Overlap

By using a window function we sort of throw away some data, when we weight the signal vector values near the ends very low. To use the data better, we perform Fourier transformation by overlapping signal input vectors. This enables us to analyse all of the data vector values almost equally, but still without introducing artifacts from abrupt "signal-discontinuities". We use 50% overlap, which is the maximum overlap without introducing redundant information.

PSD compensation from windowing

When the power spectral density values are calculated, we square the calculated Fourier transformed components.

$$\hat{G}(k) = \frac{2f_s}{N}\,|\hat{X}(k)|^2$$

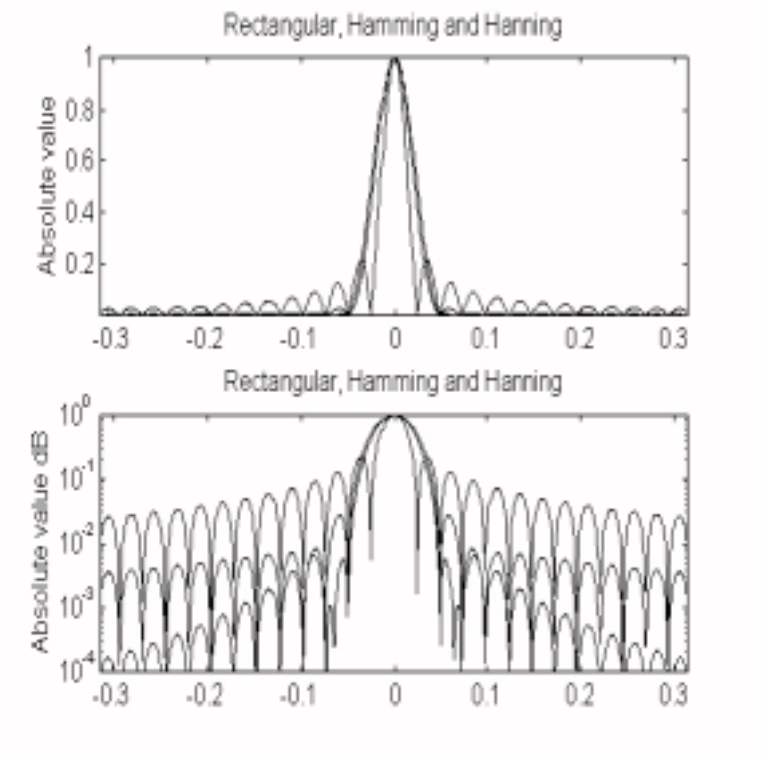


Figure 3.16: Spectra of Window Functions

The estimate of the frequency components has an effect from the windowing.

$$\begin{aligned}\hat{G}(k) &= \frac{2f_s}{N}\,|... + H(-1)X(k+1) + H(0)X(k) + H(1)X(k-1) + ...|^2 \\ &= \frac{2f_s}{N}\,(... + H(-1)^2|X(k+1)|^2 + H(0)^2|X(k)|^2 + H(1)^2X(k-1)|^2 + ...)\end{aligned}$$

The squaring of the absolute sum of $X(k)$ contains absolute values which are correlated. However, the frequency components are complex with a random phase. Carrying out the calculation of the absolute value written in the first line, reveals products of real values from different Fourier spectra. These, as well as the product from the imaginary parts, has a mean of zero. The only contribution left comes from squared values.

The total power of the Schottky signal is a sum of all powers.

$$\hat{P}_{total} = \sum_{k=1}^{N} \hat{P}(k) = \Delta f \sum_{k=1}^{N} \hat{G}(k)$$

$$\begin{aligned} &= \frac{2h}{N}\Delta f \sum_{n=1}^{N}(... + H^2(-1)|X(k+1)|^2 + H^2(0)|X(k)|^2 \\ &\quad + H^2(1)|X(k-1)|^2 + ...) \end{aligned}$$

A single frequency component appears several places in the expression above. In all a component sums up to.

$$\hat{P}(k) = X(k)(... + H^2(-1) + H^2(0) + H^2(1) + ...)$$

From this is seen that the window function is attenuating the input signal, so a compensation factor is needed for the estimate.

$$\begin{aligned} K_{Hanning} &= \sum_{n=-\infty}^{\infty} H^2(k) = \frac{1}{N}\sum_{n=1}^{N} h^2(n) \\ &= 0.3765(N = 256) \; or \; 0.3757324(N = 512) \end{aligned}$$

This is thus the attenuation for every component, compared to the ideal squared sum we have to compensate for this factor.

$$\hat{P}_{total} = \frac{2h\Delta f}{NK_{Hanning}} \sum_{n=1}^{N} X^2(k)$$

This is for a single input signal. Performing Fourier transformations of signals with x% overlap i slightly different. By this we weight the total signal with the factor.

$$K_{Hanning}(x, M) = \frac{0.40N_{averages}}{(N_{averages} + 1)(1 - x)}$$

- where $N_{averages}$ is the number of averages done.

3.6 Analysis Timing

Now we should have covered most of signal nature and the uncertainty of the calculations of its power. This section is sort of the interface between this unlimited signal frequency spectrum, to the limited bandwidth processing. It is serves as a transition to the processing.

Transverse timing

The transverse signal timing is the most important. With this signal there is a spectral zoom that has to be performed whereas the longitudinal signal treatment only needs to know the harmonic. The longitudinal signal timing, is thus left out. The transverse is explained in more details, but it is

important to underline that these parameters are only guiding. They are initial values for processing, but the system is build up so that these can be changed easily without any system knowledge, from workstations in the control room.

Whether the beam is bunched or unbunched does not change the spectrum a lot. As mentioned in section 3.3.5 there are satellites, created by the synchrotron oscillations, in stead of single spectral components. As we have a Schottky band, this does not affect our spectra that much.

The frequency spectrum of a transverse signal is sketched at figure 3.17

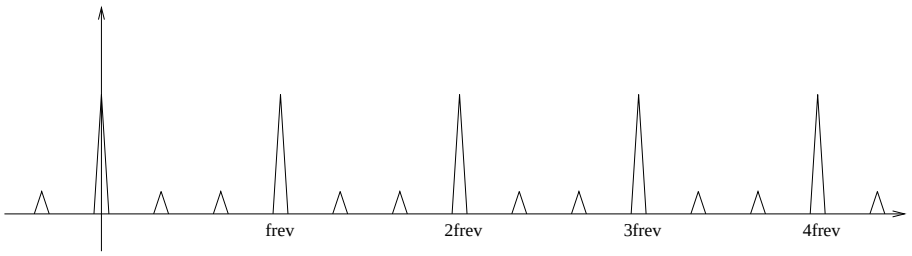


Figure 3.17: Transverse signal frequency spectrum

The harmonics of the revolution frequency is showed as being identical. This is not exactly true because we need to multiply with the envelope function, derived in section 3.2.1. However the principles, explained in the following, are not affected by this.

There are three parameters we can change to derive the frequency information of the Schottky band. These are **clock frequency, decimation ratio and local oscillator frequency**.

Clock

When the revolution frequency is constant, we sample close to the maximum clock frequency, at 40MHz. When the beam is decelerated, we sample at a multiple of the revolution frequency, $f_s = k_i f_{rev}$. The reason for this dynamic clock frequency is, that we zoom in on the Schottky band by setting a local oscillator frequency, f_{LO}. This frequency has to be a constant fraction of the sampling frequency. So if the sampling frequency was not dynamic, then the spectrum zoomed upon, would change during data acquisition as the revolution frequency changes.

The centre frequency of the Schottky band is situated at

$$f_{Schottky} = f_{rev}(m \pm q)$$

- where m is the harmonic and qf_{rev} the betatron frequency. Both m and q are constant whereas f_{rev} changes.

$$\frac{f_{LO}}{f_s} = \frac{f_{Schottky}}{f_s} = (m \pm q)\frac{f_{rev}}{f_s} = \frac{m \pm q}{k_i}$$

From the equation above, it is seen that the LO frequency can be set by only combining m,q and k_i.

The original anti-aliasing filter of the Pentek 6441 ADC has a pass band and stopband frequency of respectively 16 and 24 MHz. This is too high for our application, as we need to lower the sampling frequency more than this filter alowes. In stead another one, with a passband and stopband frequency of respectively 8 and 15 MHz, is added. The signal is damped more than 40 dB beyond this stopband frequency.

If we only alowe 40 dB damped image frequencies in the baseband and our baseband is set from DC to 7 MHz, then we have a minimal Nyquist frequency of, $7 + \frac{15-7}{2} = 11\,MHz$. So we can sample our signal with clock frequency down to 22 MHz.

Decimation Ratio

The decimation ratio reduces the bandwidth, thus enabling fast calculation of the signal Fourier transformation, see section 4.3 for details. We still need to have sufficient bandwidth, to look for our Schottky band, so we restrict this bandwidth to be 11.25% of the revolution frequency, $\frac{BW}{f_{rev}} = 0.1125$.

Unfortunately this bandwidth is affected by ripple in the passband. This ripple comes from the low pass filtering in the DRX after down mixing. The ripple becomes more significant when we approach the passband frequency. To avoid too much ripple, we only use 60% of the low pass bandwidth. This bandwidth is guaranteed to have less than ± 0.04 dB of ripple. We still perform an FFT of all the samples, but reject 40% when the transformation is done. This is equal to rejecting the samples from $0.3f_N$ to $0.7f_N$. This rather art interval in the middle of the spectrum, is because the positive frequencies of the spectrum lies between 0 and $0.5f_N$ and the negative from $0.5f_N$ and f_N. With a complex signal these negative frequencies are NOT necessarily images of the positive.

This ratio is derived beneath.

$$f_N = \frac{f_s}{4R} = \frac{k_i f_s}{4R}$$
$$\frac{BW}{f_{rev}} = 0.6 f_N$$
$$\frac{BW}{f_{rev}} = 0.6\frac{k_i}{4R}$$

Local oscillating frequency

The LO frequency is set to a presumed value of the Schottky band centre frequency. The value has to be in the frequency interval, where the head amplifiers have low noise, between 5.5 and 6.5 MHz. This area is shaded on the sketch at figure 3.18. It corresponds to the revolution frequency of 1.5 MHz and a q value of 0.4, which are realistic, values when the beam is injected.

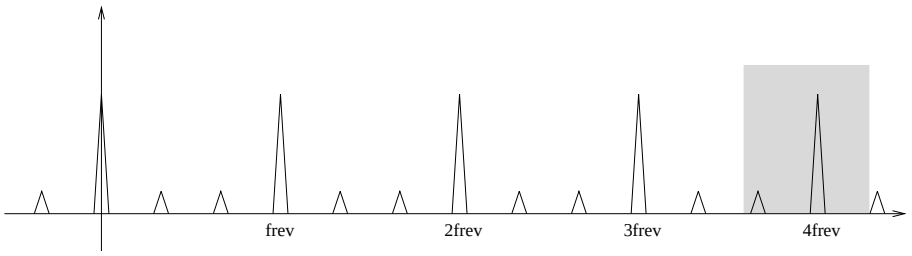


Figure 3.18: Transverse signal frequency spectrum

From this, it is seen, that we have to make a zoom upon the 4th harmonics lower Schottky band, for short 4_{minus}. This has a frequency of $(4-0.4)*f_{rev} = 5.4\,MHz$. This value is sent to the DRX, as a fraction of the sampling frequency. It is sent as 32 bit number, so the actual number sent would be:

$$\frac{5.4\,MHz}{40\,MHz}2^{33} = 1.159.641.170_{10} = 451EB852_h$$

The real q value should lie within $\pm 5.5\%$ of this guess. The q value is distinguished with steps of

$$\Delta q = \frac{0.1125}{0.6N_{fft}} = \frac{0.1875}{N_{fft}}$$

This resolution we set to be greater than 1e-3, $q \geq 1e-3$.

Constraints

We can sum up these constraints into five rules.

- The maximum sampling frequency can't be exceeded, $f_s \leq 40\,MHz$
- To avoid image frequencies in baseband we can not have a sampling frequency lower than, $f_s \geq 22\,MHz$
- The head amplifiers have the required low noise characteristics only in a limited band between $5.5\,MHz$ and $6.5\,MHz$.
- The bandwidth-revolution-frequency ratio, $\frac{BW}{f_{rev}}$, is set to be 11.25%.

- The step in q is set to be not higher than $\Delta q \geq 1e-3$

3.6.1 Parameters calculated in EXCEL sheet

The constraints mentioned above are obeyed in the EXCEL sheet shown on figure 3.19. The sheet calculates the key parameters needed, to do the digital downmixing and signal processing of the Schottky signal. Around the key parameters there are a bunch of other numbers of greater or minor interest. To get an overview of these central calculations the sheet is gone through in this section.

At the first column, we see the only given parameter for this sheet, the variation of the momentum, p. This momentum is linear between the two columns, "p from" and "p to". Each row is a time-slice in which all hardware and software settings are constant. In a popular sense, this can be thought of as a "car-gear" that doesn't need to be changed, even though the speed changes a little bit. In our system this means that neither the setup of the DRX nor the FFT parameters changes during this time, even though the sample frequency changes a little. The duration of the time slice is depended on the amplifiers. The zoom has to be done within a specified range restricted by the amplifier characteristics.

The different states of the beam are divided by shaded horizontal bars. For the Schottky measurement there are only two important states, these are when the beam has a constant revolution frequency, hence constant momentum, or varying revolution frequency. The different momentum levels are graphically shown in time on figure 2.9 on page 28.

In the EXCEL sheet there are some vertical columns shaded light gray, these are the key parameters from which almost all the other columns are derived. There are however columns concerning the fast Fourier transformation, which are decided independently.

As appendix on Appendix A all of the column names are explained and possibly added a formula for the calculation of the column. The information should be complete, thus reproduction of the Excel sheet can be done from this.

Transverse sample clock and Schottky band harmonics

Momentum		Revolution frequency		Sampling Frequency			Decimation factor	q bandwidth	Schottky band chosen		Local oscillator frequency			time of measurement				q resolution		Sampling rates		Angle of envelope	
p from (GeV/c)	p to (GeV/c)	f_rev from (MHz)	f_rev to (MHz)	K_i (integer)	f_s from (MHz)	f_s to (MHz)	R (integer)	BWf_rev h ()	harmonic (integer)	sign of q ()	LO from (MHz)	LO to (MHz)	F2^32 ()	N_av ()	N_FFT ()	meas.ti me* (msec)	# of meas. ()	delta_qRIN (in 1e-3)	total period (msec)	Complex sample rate (highest)	Complex sample rate (lowest)	START: angle of exp.funct. envelope at f_s/2 [degrees]	END: angle of exp.funct. envelope at f_s/2 [degrees]
constant:		1.590	1.590		40.000	40.000																	
3.592	3.592	1.590	1.590		40.000	40.000	32	0.1179	4	minus	5.724		0.143	16	512	13.9	1436	0.38	20000	312500		6.84	8.09
decal. to		1.590	1.590																3500				
3.592	2.359	1.590	1.527	15	23.850	22.905	20	0.1125	4	minus	5.724	5.497	0.240	16	512	15.2	178	0.37	2714	298125	286313	6.84	8.09
2.359	2.001	1.527	1.488	15	22.905	22.320	20	0.1125	4	plus	6.719	6.547	0.233	16	512	15.6	50	0.37	786	286313	279000	7.12	8.42
constant:		1.488	1.488		40.000	40.000											228		15000				
2.001	2.001	1.488	1.488				36	0.1120	4	plus	6.547		0.164	16	512	15.7	957	0.36	15000	277778	277778	7.31	8.64
decal. to		1.488	1.488																3500				
2.001	1.099	1.488	1.250	24	35.712	30.000	32	0.1125	4	plus	6.547	5.500	0.183	16	512	18.6	99	0.37	1956	279000	234375	7.31	8.64
1.099	0.741	1.250	1.019	27	33.750	27.513	36	0.1125	5	plus	6.750	5.503	0.200	8	512	12.1	61	0.37	736	234375	191063	8.70	10.29
0.741	0.575	1.019	0.859	27	27.513	23.193	36	0.1125	6	plus	6.522	5.498	0.237	8	512	14.3	23	0.37	342	191063	161063	10.68	12.62
0.575	0.475	0.859	0.743	42	36.078	31.206	56	0.1125	7	plus	6.357	5.498	0.176	8	512	16.5	12	0.37	205	161063	139313	12.66	14.97
0.475	0.397	0.743	0.640	42	31.206	26.880	56	0.1125	9	minus	6.390	5.504	0.205	8	512	19.2	8	0.37	162	139313	120000	14.64	17.30
0.397	0.318	0.640	0.528	42	26.880	22.176	56	0.1125	10	plus	6.656	5.491	0.248	6	512	19.1	8	0.37	161	120000	99000	17.00	20.09
0.318	0.300	0.528	0.501	45	23.760	22.545	60	0.1125	12	minus	6.125	5.812	0.238	6	512	19.1	1	0.37	37	99000	93938	20.60	24.35
constant:		0.501	0.501														212		6000				
0.300	0.300	0.501	0.501		40.000	40.000	104	0.1152	12	minus	5.812		0.145	16	512	45.3	132	0.37	6000	96154	96154	21.71	25.66
decal. to		0.501	0.501																2000				
0.300	0.259	0.501	0.437	78	39.078	34.086	104	0.1125	13	minus	6.313	5.506	0.162	12	256	20.3	20	0.73	414	93938	81938	21.71	25.66
0.259	0.221	0.437	0.377	78	34.086	29.406	104	0.1125	15	minus	6.390	5.504	0.187	10	256	19.9	18	0.73	376	81938	70688	24.90	29.42
0.221	0.193	0.377	0.331	78	29.406	25.818	104	0.1125	17	minus	6.258	5.495	0.213	8	256	18.6	15	0.73	281	70888	62063	28.86	34.10
0.193	0.165	0.331	0.284	78	25.818	22.152	104	0.1125	19	plus	6.421	5.510	0.249	8	256	21.6	13	0.73	283	62063	53250	32.87	38.84
0.165	0.140	0.284	0.243	87	24.708	21.141	116	0.1125	23	minus	6.418	5.492	0.260	8	256	25.3	9	0.73	243	53250	45563	38.31	45.27
0.140	0.119	0.243	0.207	108	26.244	22.356	144	0.1125	27	minus	6.464	5.506	0.246	8	256	29.7	7	0.73	211	45563	38813	44.77	52.91
0.119	0.100	0.207	0.174	129	26.703	22.446	172	0.1125	31	plus	6.500	5.464	0.243	8	256	35.3	5	0.73	192	38813	32625	52.56	62.11
constant:		0.174	0.174														87		1000				
0.100	0.100	0.174	0.174		40.000	40.000	306	0.1127	34	plus	5.996		0.150	1	256	7.8	128	0.73	1000	32680	32680	62.52	73.89

constants:
alfa = 0.6
q = 0.4

Figure 3.19: Measurement state data

3.7 Beam transfer function(BTF)

Our aim of processing is to be able to analyse the Schottky noise signals. However it might happen, that these signals are too weak for analysis. In such a case there is another possibility for tune measurement, namely the beam transfer function calculation. This is an older and simpler method of measuring. The method is only mentioned here, as the report has its main focus on the Schottky signal processing.

Calculation of the beam transfer function, involves disturbance of the beam. Basically, the beam is kicked a little bit and the response is measured by the pick-ups. What happens physically is that the oscillations becomes larger and thereby the pick-up signals as well. The snapshot of the beam on figure 2.7 is at the time of kick displaced vertically. Each particle is still moving around the origin, the shape changes to a "snail-like" trajectory around the origin. As the area, also known as the normalised emittance, stays the same, the fundamental theorem of Louville still holds.

The calculation of the BTF simply has to signal inputs, the kicker signal and the detected beam signal. These are Fourier transformed and the ratio between them are found.

$$BTF(f) = \frac{Detected(f)}{Kicked(f)}$$

Chapter 4

Hardware

This section begins with an overall description of the designed system at block level. Each an every block is shortly introduced and the functionality in the system mentioned. Blocks, such as the ADC and DRX boards are described slightly more detailed than the others. More profound descriptions follows in the sections describing the downconverter and the digital signal processor chip. A range of features is introduced in order to prepare the reader for the later sections describing the written software, using these features.

4.1 Overall system description

The control of a synchrotron is a delicate matter. First of all there is a great need of accuracy and secondly it has to happen at a very specific time. Some tasks has to be controlled down to 1 millisecond, so timing is an important parameter. The system, to be controlled, is shown on figure 1.3 on page 12. Note that building are situated in the middle of the PS synchrotron, so that there is the same signal delay to all cells of the lattice. Underlining the importance of high accuracy timing. The AD is situated to the left, as a rectangular building at the left with an oval-like lattice inside.

The AD control system, of which the embedded Schottky analysis system is connected to, is quite large. There are a lot of blocks that works in parallel controlled by central control units, the device stub controllers(DSCs). A block diagram is shown on figure 4.1. The description is first divided into the blocks, shown on the figure and later the measurement procedure of the system, as a whole, is described.

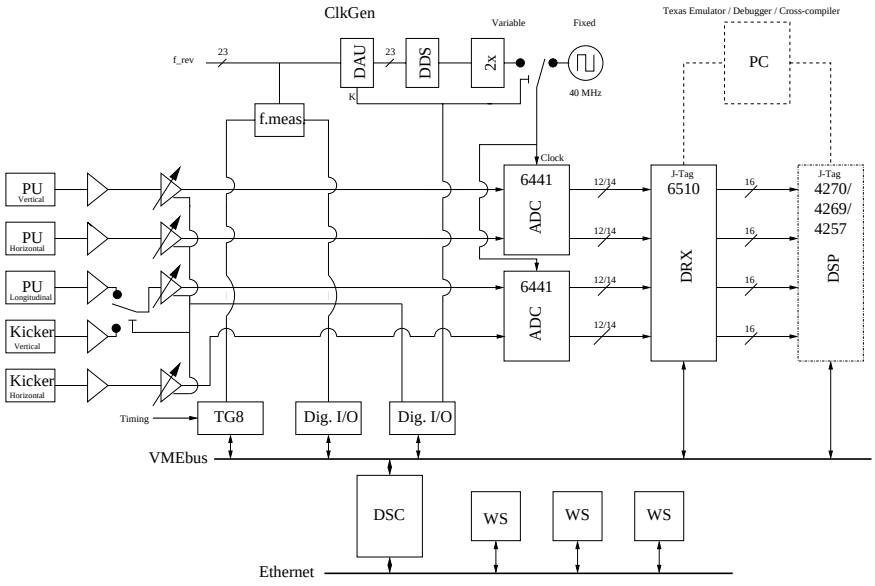


Figure 4.1: Block diagram of acquisition and processing system

4.1.1 Description of blocks

Pick-ups(PU) and amplifiers

There are three pick-ups made, for the beam detection. A transverse horizontal, a transverse vertical and a longitudinal pick-up. The induced currents on the pick-ups surfaces are quite weak and they are amplified by head amplifiers sitting on the pick-ups them selves. These amplifiers are a delicate matter. The noise contribution of these are made as low as possible. The head amplifiers on transverse pick-ups has a noise level between $0.3 - 0.5\,pA/\sqrt{Hz}$ in frequency interval between $5.5 - 6.5MHz$. The longitudinal signal analysis needs a larger bandwidth, so three amplifier gain modes are used. The power spectral density noise is either $1.6\,pA/\sqrt{Hz}$ or $15\,pA/\sqrt{Hz}$ depending on the gain mode.

After the head amplifiers, there is another amplifier/attenuator. It is controlled by software, in order to use the maximum dynamic range of the ADC. The amplifier/attenuator has a gain from -17 to +28 dB in 16 steps. 4 bit codes the amplification level, so the LSB is equal to $\frac{28--17}{2^4-1} = 3dB$.

The amplifier also has an anti-aliasing filter, which matches the requirements of the quantisation. Two filters are present, the existing Pentek anti-aliasing filter and the added filter having lower bandpass characteristics. The switching to and from the additional filter is under software control.

Timing Module(TG8)

For the system timing, a CERN TG8 timing module is used. This module has 8 output channels and can receive an input request of an interrupt via the VME bus.

Digital Arithmetic Unit(DAU)

As incoming signal to our system, we have the revolution frequency represented in 23 bits. This frequency has to be multiplied by the k_i factor to obtain the sampling frequency information. This DAU carries out this multiplication.

Direct Digital synthesiser(DDS)

A DDS module converts the sampling frequency information into a train of pulses at half the sampling frequency. The module can only generate a pulse train of maximum 20 MHz.

Doubler (2x)

As we need a pulse train at a maximum frequency of 40 MHz, we need to double the the frequency and this is done by a doubler module.

Fixed clock

A fixed digital pulse train of 40 MHz is produced by a crystal oscillator module.

Digital I/O

The digital I/O modules serve as interface of data from the VMEbus to modules not attached to this bus.

Pentek 6441 Analog to Digital Converter(ADC)

Figure 4.2: VME crate embedded Pentek 6441 ADC card

As converter we use a an VME crate embedded Pentek 6441 analog to digital converter which quantifies in 12 bits at 40 MHz. The specifications for this ADC is written in appendix B. They are directly taken from the

manual, which is currently[1] available on the Internet at [19]. The card is shown on figure 4.2

The internal block diagram of the ADC is shown at figure 4.3

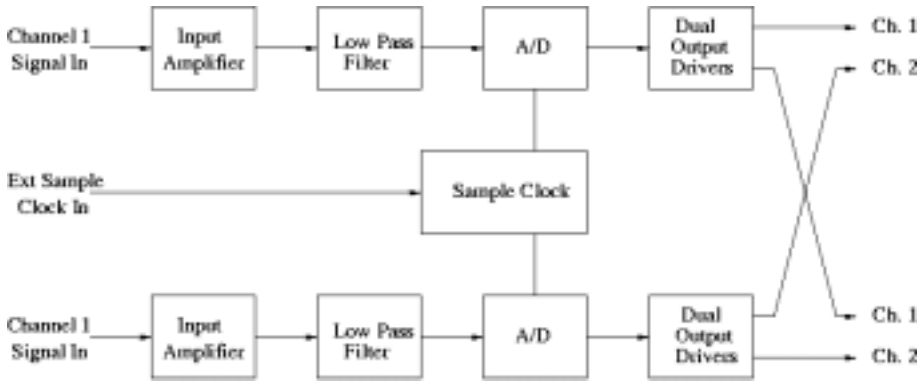


Figure 4.3: Block diagram the ADC

Pentek Digital Receiver 6510 (DRX)

Figure 4.4: VME crate embedded PENTEK 6510 digital receiver card

As receiver of the digitized signal the Pentek 6510 Multiband Digital Receiver MIX module is used, see card on figure 4.4. The digitised data is received at a sample rate of up to $40\,[MHz]$. The data is reduced in the receiver to minimise the calculation load for the subsequent real time FFT analysis. The specifications are written in appendix C.

With the digital mixer, the interesting frequency band is down converted to a band at low frequencies. This band is lowpass filtered and decimated. The principle is that the band of interest, at first, takes up very little of the whole band. Thus making a zoom by just removing all other frequency information leaves you with very few samples. Then details can be inspected after a subsequent Fourier transformation of the reduced data. The principle is described in section 5.2.1 on page 84.

At the board there is as well a Texas Instruments DSP, TMS320C40. This is a central component for this project and will be discussed a lot more

[1] September 1998

profound later in this report, see section 4.4.

Similar to the ADC there is, at the moment, available manuals and specifications for the 6510 DRX on the WEB at [19].

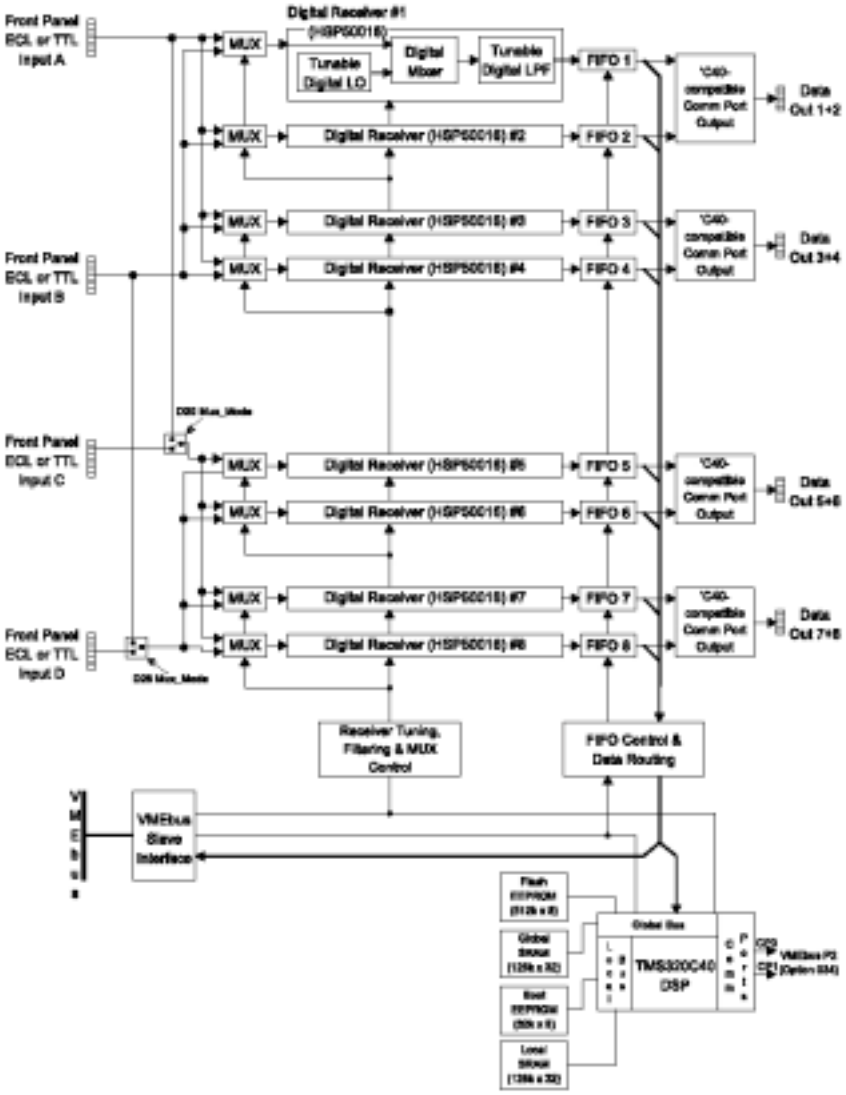


Figure 4.5: Block diagram of Digital Receiver

Digital Signal Processor (DSP)

The digital signal processor board on the block diagram is not included in the current system version. This is an option for further development, when the cheaper version has shown its worth. This is why the board is shown with punctured lines.

A version with DSP board will include a board with multiple processors. The boards considered was the Pentek 4269, 4270 and 4257 DSP boards all with either 2 or 4 TMS320C40's. Information about these boards can be found at [19].

Device Stub Controller (DSC)

The backbone of the control system is the **Device Stub Controller, DCS**. The DSC has a CES RI08062 CPU (PowerPC) and are embedded in VME crates. There are more than 100 of these DSCs in the PS complex. The DCS controls every device attached to the common bus, in terms of requests, timing, handshaking, addressing etc. The operating system of this DSC is LynxOS 2.5. It is a unix based real-time operating system. As all other real-time operating systems, LynxOS works on an interrupt basis, with interrupt rates down to 5 milliseconds

The DSC is connected to Ethernet through an interface. It is also connected to a VME bus. The VME bus system is a 160 PIN bus connection with a transfer rate of 160 $Mbyte/sec$.

Work stations(WSs)

The **workstation** in the PS network are unix based terminals and PCs. The workstations have IBM RS/6000 CPUs and they're using the Unix AIX operating systems, which is the IBM made unix system. The PCs are using Windows 95/NICE and are communicating with the control system via an X-console. They are both connected to Ethernet, which runs the communication protocol TCP/IP[2].

4.2 Measurement procedure

All parameters of the measurements are controlled by work stations. They give a measurement request to the device stub controller, DSC, via Ethernet. This request contains data of everything concerning system measurement setup. When data is received by the DSC, it takes full control of the measurement. Time interrupts are requested of the TG8 and all of the system is set up via the VME bus. The following things are set up before the data acquisition can begin.

- levels of signal amplification, -17 - +28 dB
- choice of measurement, Schottky or BTF
- choice of clock source, fixed or variable

[2]Transfer Control Protocol/Internet Protocol

- setup of variable clock factor, K
- setup of DRX, 8 control words of 40 bits to each receiver chip
- setup of DRX data buffer
- setup of a DSP processing request

When the requested time interrupt occurs, the DRX buffers are enabled thus collecting downsamled data for processing. When the buffers are half full, they interrupt the DSP so that the DSP can empty data buffer and process data. When a sufficient number of data is collected, the DSP disables the data buffer of the DRX. When the DSP has finished the calculations on data, the DSC is informed. The DSC can then take the calculated data, from global shared memory, and deliver it to work stations, where it is presented.

This is the main lines of the overall procedure. The specific internal publication defining this, is not finished at the time of the writing of this report. The structure is however clear and the software modules are written, so that they can be changed whenever the final specifications are ready. An early draft of the publication is made by the writer of this report, but it needs modifications and approval of a system aquianted. The definitions are bound to be ready towards the end of this year, 1998.

4.3 Harris HSP50016 DRX chip

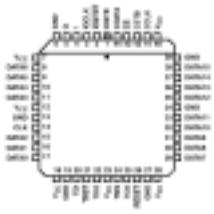


Figure 4.6: The Harris HSP50016 downconverter chip

The HSP50016 Harris downconverter chip basically only has two important features. The down conversion to baseband and reduction of bandwidth. This is the same as watching the signal spectrum through a window of optional width, at an optional spot. The frequency spectrum range, goes up to 75 MHz and in this range the chip can produce output with a bandwidth of between $\sim$ 600 Hz to $\sim$ 1.2 MHz. At the maximum data rate, the baseband for down conversion can be set at 0.01 Hz of precision.

The down-converter chip is the fastest component in the system developed. The Pentek 6510 Digital Receiver contain 8 of those 48 pin digital down converter chips. This is what enables us to perform spectral analysis, at such data acquisition rates, hence we acquire close to 2 GSPS[3]. The digital receiver, also denoted DRX, can acquire and process 16 bit of input data in up to 75 MHz, thus a possible data rate of 1.2 GSPS[4] per chip. This is not possible to process with any existing DSP, so without the down-mixing this spectral analysis could not meet the real time constraints.

What the Harris chip is doing is basically shown on figure 4.7. The principle of downmixing is explained in section 5.2

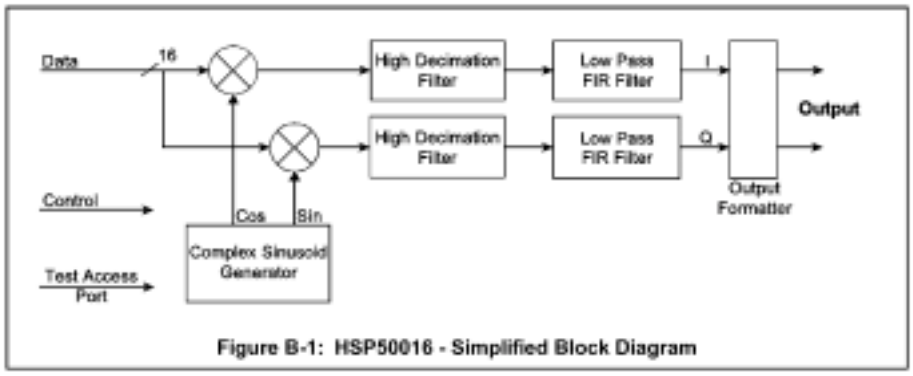


Figure 4.7: Simple internal block diagram of the Harris chip

4.3.1 DRX functionality

The 16 bit input signal is multiplied by two 16 bit cosine value at the sample rate one for each channel (real and imaginary). In our system, we only have a 12-bit ADC, so the LSBs are set to ground. The multiplied value is digitally filtered and decimated by a high decimating filter, HDF. It is called a filter because it is really just an ordinary FIR[5] filter, but with unit values, thus an integrator followed by a decimator.

To conserve the 16-bit of resolution, the products are always shifted up to MSB. Calculations are always performed in registers with more than 16 bits of resolution. A scaling multiplier is added to the HDF output to compensate for attenuation of decimation rates which are not powers of two.

The filter response of the HDF square-type filter is a sinc function, which has its first zero value at $\frac{f_s}{R}$, where R is the HDF decimation rate. The

[3]Giga Samples Per Second
[4]Giga Samples Per Seconds
[5]Finite Impulse Response

signal passes through another 121-tap FIR anti-aliasing filter that improves the filter response in pass and stopband. This filter is also of a decimation filter type, but with a fixed decimation of 4. Then only a fairly flat part of the HDF filter response is used and this following filter compensates for the declination on the HDF filter response characteristics. This composite filter has a bandpass ripple of less than 0.04 dB and a stop band attenuation of more than 104 dB, see figure 4.8.

Because of this composite filter only 60% of the down converted bandwidth can be used, the rest is too attenuated. The usable bandwidth, from the signal output, is then $0.6\frac{f_s}{4R}$ and the value setup in the Harris chip is the decimation factor[6], which is $R-1$.

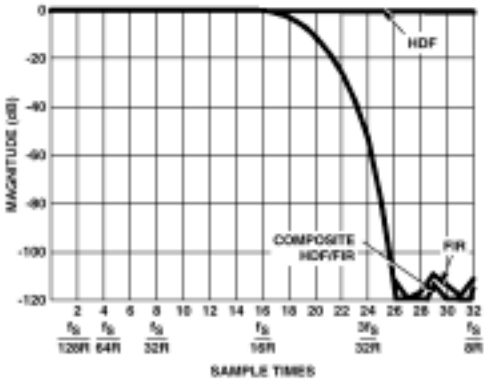


Figure 4.8: HDF/FIR filter response characteristics for R=16 (minimum value)

The Harris chip can decimate an incoming signal by factors from 64 (2^6) to 131.072 (2^{17}), in order to reduce the bandwidth. It is important to underline that the baseband is digitally transformed without any loss of baseband information. The data rate from the Harris chip is $\frac{f_s}{4R}$, where f_s is the clock frequency and R the HDF decimation rate.

4.3.2 Formatter

The down converted signal passes through an output formatter. This formatter enables the programmer to chose the format of the down-mixed signal. First of all the timing to the connected memory chip, has to be defined. There are several pins on which the data can poor out from and in different ways. Further more the timing parameters has to be set. This is all defined by Pentek which designed the board and thus decides the Harris chip output to FIFO data buffer memory interface.

We can chose how the numbers should be stored in these FIFOs. There is an option of storing it as:

[6]Note! the difference between decimation rate(R) and decimation factor(R-1)

- floating point
- binary offset
- 2's complement

The floating point format is stored similar to the TI floating point format, which has an implied bit and thus gains a bit of resolution, see section 4.4.5 on page 75. The word length can be set to 16, 24, 32 or 38 bits. The last option represents full representation of last internal register calculations. The others are symmetrically rounded off to LSB, which means that the 38 bit binary representation are simply cut off. A saturation option can be set so in case of saturation only the maximum value, positive or negative, is put at the output. There is another option which enables surveillance of a defined threshold value. The threshold values can be 1/8, 2/8, ... 7/8, 8/8 of the maximum range. A register calculates the number of times this value is exceeded or an interrupt to the on-board DSP, can be ordered.

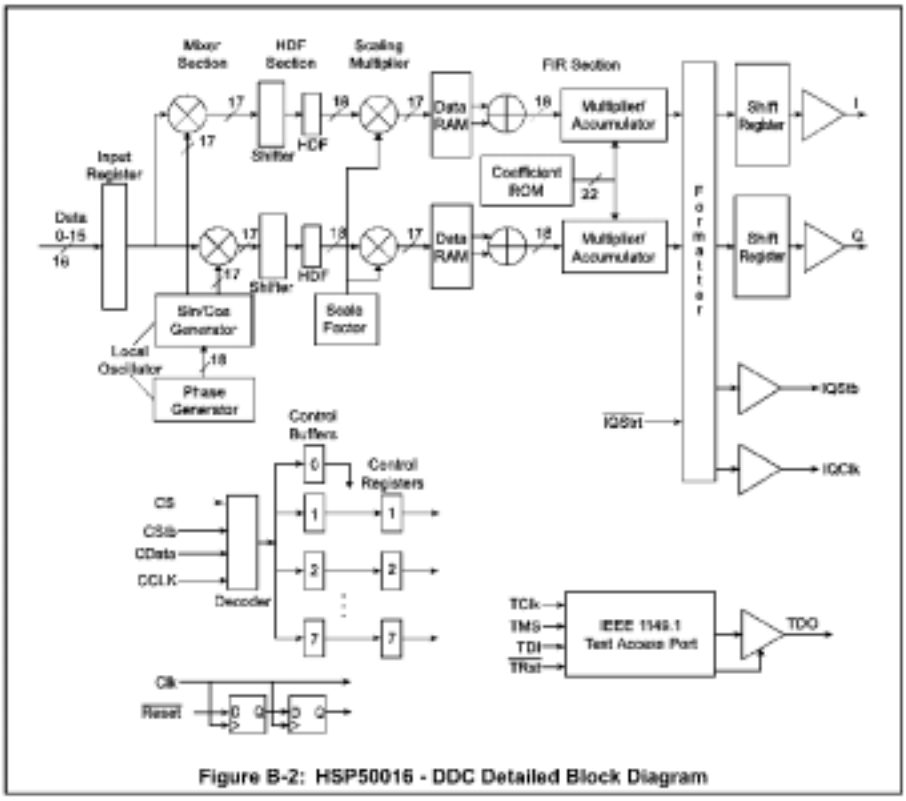


Figure 4.9: Extended internal block diagram of the Harris chip

4.3.3 DRX setup

The Harris chip is setup by 8 control words each of 40 bits. The most important information is the local oscillator frequency and the decimation rate. Other setup parameters are variable flag indications, interrupt settings, signal conditioning, signal timing settings, etc. The control words are shifted in 4 by 4 to the Harris chips. This serial process can be done either by the DSP via local bus or through the VME bus by external hardware. In our system the DRX is setup by the C40 via local bus.

On the Pentek DRX board, there is some firmware in on-board flash ROM. This firmware is used to setup the 8 on-board Harris chips. Through company and additional software the control words are created. There are certain bits in the control word that we do not have the liberty to control. For example we do not know how they connected the FIFO memory to the Harris chip output, so timing settings is not for us to set. In principle we can set these bits as well, but we chose to use the company software to make this DRX setup right. The company software is written in C and was delivered free of charge. Minor errors in these programs was found and they are corrected in the version used.

More information about the Harris HSP50016 digital downconverter chip can be found from [17].

4.4 TMS320C40 architecture

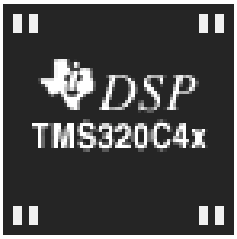


Figure 4.10: The Texas Instruments TMS320C40 digital signal processing chip

The TMS320C40 is a digital signal processor from Texas Instruments(TI), which runs at a maximum clock frequency of 50 MHz. The maximum performance is 275 MOPS[7]. The structure description of this DSP, is divided into sections describing the blocks. A block diagram from the TI data sheet[8] is shown in appendix D. The descriptions are mainly based on man-

[7]Mega Operations Per Second
[8]datasheet available on the WEB, http://www.ti.com

uals from TI [11, 12, 13, 14] and course material from the TI technical training course at Cranfield University(UK) [15].

4.4.1 Pipelining

Ordinary processors handles a command in one or more clock cycles, depending on the nature of the command. There are always different tasks involved in a single command and a normal processor carries them out, more or less, in series. The 'C40 processor works differently. It uses a pipeline architecture which is a parallel architecture. This means that every command goes through 4 stages in the processor at different ticks of the clock. These stages are in order of arrival fetch, decode, read and execute. So when an assembly command is about to be executed an instruction, three instructions ahead, an instructions is about to be decoded in 'C40 registers.

If the execution is a command that changes the successive read-in of commands then this pipeline is flushed. This means that preparation of former commands is wasted and the processor restarts the pipeline. This means that the programmer does not have to bather about the pipeline because the processor it self, takes care of the pipeline.

However ! If the aim is to produce fast software code, then the software should profit from this high speed architecture. The assembly command set contains commands which wraps the pipeline around, so that the flushing is minimised. An example of this is the important repeat block command (RPTB). This command is used when a block of commands are carried out, a number of consecutive times. Then the fetch automatically returns to the start of the block, in stead of reading further on passed the end of the block. This way the block fills the pipeline with perfect overlap enabling maximum processing speed.

Another command set option is to make a command delayed (D). This means that when the command is at the execution state, it will wait three clock until execution. This way the pipeline do not need to be flushed and another case of perfect overlap can be obtained.

The TI C-compiler has optimisation options that modifies the assembly or C code so that, among others, this pipeline is used to gain speed. The optimiser is however not yet as fast as a good programmer can do, with a little bit of creativity. This is why the core of a the software should be written in assembly, if the processing time is an issue.

The processor undergoes a pipeline conflict if the rules of commands are not followed right. There are three kinds of conflicts: Branch, Register and Memory conflicts, these are all a cause of bad placement of commands, but does not always lead to an error when assembled. Such conflicts has occurred in the code written, but are now corrected. The symptom of such an error is that the execution of single instructions is performed in

an unpredictable highly peculiar way. The conflicts are described in [11] section 8.2.

4.4.2 Addressing modes

The 'C40 processor has a 32-bit address bus to address memory, thus 4 Giga locations. Of registers the processor has 32 registers with software access. The addressing can be categorised in four categories, immediate, direct, register and indirect.

- Immediate addressing is when the command it self holds the value of a 16 bit numerical constant. Assembly example: LDI 02H,R0
- Direct addressing uses a Data Page Pointer (DP register) to access a memory cell. It only stores the offset from the Data Page Pointer base. That way the DP register holds a 16 bit Data Page Pointer and the command word only 16 bit. Like this 65k memory cells can be accessed without changing the DP register. Assembly example: LDI @x,R0
- Register addressing is straightforward like every other processor. Assembly example: LDI R1,R4
- Indirect addressing uses the auxiliary registers (AR0-AR7) to access a memory cell. This addressing type is quite frequent when dealing with data series because the auxiliary registers can be updated in the same command. It is thus a fast way of treating all of the data cells in a systematic way. Assembly example: LDI *AR1,R0

Moreover there are several modes of addressing used for specific applications. One of them is bitreversed addressing which is used for storing FFT results, which by nature is located in a bitreversed order. An important constraint to these instructions, is that the addressing has to be done in a memory segment with zero offset. Another mode is the circular addressing, used with data structures that are to be run through in a cyclic way.

4.4.3 Registers

The 'C40 has 32 registers accessible by the programmer. The most frequently used registers are the 12 40-bit extended precision registers (R0-R11) and the 8 32-bit auxiliary registers(AR0-AR7). A list of registers are shown beneath, every one of them 32 bit long, except the extended precision registers.

- R0-R7, extended precision registers
- AR0-AR7, auxiliary registers

- DP, Data Page Pointer register
- IR0-IR1, Index registers
- BK, Block size register
- SP, Stack pointer register
- ST, Status register
- DIE, DMA coprocessor interrupt enable register
- IIE, Internal interrupt enable register
- IIF, IIOF flag register
- RS, Repeat start address register
- RE, Repeat end address
- RC, Repeat counter register

When using registers in assembly instructions, it has to be specified whether it has to be saved as an absolute number (integer) or floating point. The assembler do alowe inconsistence and the consequence is scrambled data. The extended precision registers are different from the rest. When they store an integer, it is done in the lowest 32 bit. When it is a floating point number it uses the entire 40 bits. When a 40-bit floating point number is stored in 32-bit memory, it can be done in two ways. Either there is a loss of resolution corresponding to 8 bits or it is stored in two memory cells. The last option is used when a extended precision register is pushed onto the stack. By using successively PUSH Ri and PUSHF Ri all bits are conserved.

The floating point representation reserves 8 bits for the exponent and the rest for the mantissa. The rest is thus either 32 or 24 bits, containing the sign and fraction of the floating point number.

The auxiliary registers has two separate arithmetic units for calculations. This means that in parallel with normal instruction, they can be incremented or decremented in different ways. Signs($+/-$) put before an auxiliary register means that the increment/decrement is to be performed before it is used in the instructions. Double signs($++/--$) means that the register is modified afterwards. The parenthesis, after the auxiliary register, indicates the increment/decrement factor[9] which can be 0,1 or content of either IR0 or IR1. If bitreversed addressing is wanted then a B is simply added at the end.

Some examples of legal register modifications: AR4, -AR0, ++AR1(IR1), AR2–(IR0)B

[9]By default 1, if there is no parenthesis

4.4.4 Memory map

As mentioned before there is 32 bits for addressing, thus enabling addressing of 4 Giga words of memory. The MSB bit of the address is the ROM enable bit (ROMEN) if this is set to 1 then the address space is limited to the 1M reserved memory. Normally the ROMEN is set to 0 enabling 2 giga of address space. Apart from the first 3 Mega of special purpose memory, memory mapped registers, memory mapped I/O etc. this is divided between external local(2G-3M) and global(2G) bus memory.

The internal memory consist of two blocks of 1k fast dual access RAM and memory mapped access' to peripheral registers and ports.

4.4.5 TI floating point

Normal IEEE standard floating point numbers consist of an exponent and a mantissa. The exponent is a 2's complement binary number indication the position of the binary point. The MSB of the mantissa indicates the sign of the fraction, 0 for positive and 1 negative negative. This is because the fraction contains a 2's complement number with the only difference that it is shifted until the MSB[10] and the NSB[11] are different. The exponent hold the number of shifts, positive or negative. The fraction of a positive number will then lie between 0.5 and 1 and of a negative between -1 and -0.5.

As a fraction of a positive number always starts with 01 and a negative 10, it holds redundant information. TI has decided to throw away the second bit leaving only the sign and holding in mind that they have to inject an "implied bit" when converting. By this one bit is gained compared to normal IEEE format.

An example of conversion to IEEE and TI floating point formats is shown beneath.

The number 43.0 as absolute binary number is represented as:

$$..00101011.0000_2..$$

,we have to shift the point 6 times to obtain a true fraction, leaving the exponent as 00000100_2 and the number[12] as:

$$0\mathbf{0.101}0110000_2$$

In an IEEE floating point format, this is represented in hexadecimal as:

$$06560000_h$$

[10] Most significant bit
[11] Next significant bit
[12] first fractional hexadecimal cifre boldfaced

In the TI format we shift until we have a true fraction, neglecting the first 1. This leaves us with an exponent of 00000110_2 and a number of:

$$0\mathbf{01.010}110000_2$$

Then we throw away the 1 before the binary point and gets a hexadecimal representation of:

$$052c0000_h$$

From this can be seen that hexadecimal comparison is quite difficult to do, by heart. Therefore for those equipped with a HP48G/GX, the author of this report has made a short program, that converts hexadecimal represented TI floating point numbers, into ordinary floating point numbers. This is a quite useful tool when debugging, because not registers in the debugging tool can be shown in floating point format. The program is enclosed in appendix E.

4.4.6 Assembly instruction set

The 'C40 assembly instruction set supports 135 instructions. Most of them known from the basic assembly instruction set. Basically there is only four differences in the instruction set compared to basic assembly.

- As already mentioned before there is a pipeline structure that introduces more instructions. The two categories of instruction are delayed and repeated instructions.
- As there are different bit representation in registers, there is instructions which compensates for lost resolution.
- The 'C40 uses a TI floating point format different from normal standard. There are instruction converting this format to the conventional used.
- As the pipeline enables parallel tasks, then a certain combination of instructions can be executed at the same time.

The last item in list, concerning parallel instruction, needs to be explained a little bit more. As the 'C40 has separate multiplier, ALU[13], it can perform parallel arithmetic. There is also two Auxiliary Register Arithmetic Units (ARAUs) so addressing can be performed in parallel with the other units. This is to some extend four parallel CPUs, so in one single clock cycle four tasks can be performed. An example of this could be:

[13] Arithmetic and Logic Unit

```
    MPYF3   *AR0++(1),*AR2--(IR0),R0
||  SUBF3   R1,R2,R3
```

This single command[14] performs the following tasks.

- Content of address held in AR0 is multiplied with the content of the address held in AR2 and stored in R0.
- The content of R2 is subtracted by the content of R1 and stored in R3
- After the execution, the AR2 register is incremented by 1
- After the execution, the AR0 register is decremented by the content of IR0

There is some restrictions to the use of parallel instructions. Only two types of addressing can be used, register and indirect addressing. Moreover, it is not always any combination and any registers that can be used. In the above mentioned example, it is restricted that the first two addressing types in the two commands has to contain two registers and two indirect addressing. The last one of the MPYF has to be either R0 or R1 and the last one of SUBF either R2 or R3. These restrictions makes it necessary to puzzle with the order, the choice of register and command types in order to obtain a short compact code. What is gained however is sometimes quite impressive, an improved speed of 100-200%, is not unusual.

4.4.7 DMA data transfer

The 'C40 has a separate 6 channel DMA[15] coprocessor that can take care of the data transfers, without occupying processing time. Every channel has memory mapped DMA registers, where the transfer request can be stored. The 'C40 has two modes of transfer, unified and split. Only the unified mode is used and therefore only this explained in this report. A data transfer request to a single channel contains the following.

- DMA channel control register
- Source address
- Source address index (A signed number indication the increment/decrement steps of the source data)
- Transfer counter (The number of words to be transfered)
- Destination address

[14] || means that the commands are to be merged into one
[15] Direct Memory Access

- Destination address index (A signed number indication the increment/decrement steps of the destination data)
- Link pointer
- Auxiliary transfer counter
- Auxiliary link pointer

When a transfer is requested, then the request is put in DMA registers[16] and an interrupt flag is set[17]. If an interrupt is wanted, when the DMA has finished the transfer, then another flag has to be set[18]. The DMA can auto-initialise if the link pointer is pointing to another request anywhere in the memory space. This way is is possible to have a chain of transfers happening in parallel with the CPU processing.

A single transfer request requires thus only transfer of 10 words, where the control word is the last one that initiates the transfer. Such a transfer request can be minimised to occupy only 5 clock cycles and as an assembly function with push and pops such a request occupies only approximately $0.4\,\mu sec$.

The channels are prioritised, the zeroth channel has the highest priority. If demands are made to several channels, then they can share DMA processing time. These settings and priorities are among others, set in the control word of each request.

[16] channel 0 'C40 address 0x10 00A0 - 0x10 00A8

[17] The Transfer Counter Interrupt (TCINT) flag (20th bit) in the DMA channel control register

[18] The transfer Counter Interrupt Control (TCC) bit (18th bit) in the DMA channel control register

Chapter 5

Aspects of processing

This chapter covers some of the different aspects of the processing involved. They are put together in this chapter, because they don't fit in nicely elsewhere in this report. Most aspects are known to people familiar with signal processing, so they might want to skip this chapter. However, the importance of these techniques and aspects to this project are significant. To obtain a complete picture of the processing done, these aspects should be familiar.

The chapter starts out with a description of the signal quantisation. Calculations on signal noise contributions, as cause of this quantisation, is gone through. Following is a description of the complex downmixing of the quantified signal. The principle is described and derived from mathematical point of view. The last section covers the Fourier transformation, especially the fast Fourier transform algorithm (FFT). It is shown how the fast Fourier transform algorithm is derived from the normal Fourier algorithm. Basically, it is the 3 important stages the signal goes through, before more specific software, derives the last parameters.

5.1 Quantisation

The quantisation of the incoming analog signal, is done by a Pentek 6441 board. This board uses the Analog Devices AD9042 ADC chip. More information about this chip, can be found on Analog Devices homepage [20]

When a signal is quantified to a certain number of bits, an error is added to the signal. This error is zero in mean and is thus distributed rectangular between $-\Delta_{max}$ and $+\Delta_{max}$. The maximum error is half the interval of which the voltage range is divided into.

$$\Delta_{max} = \frac{1}{2}LSB = \frac{1}{2}\frac{2}{2^{12}-1} = \frac{1}{4095} = 0.244\,[mV]$$

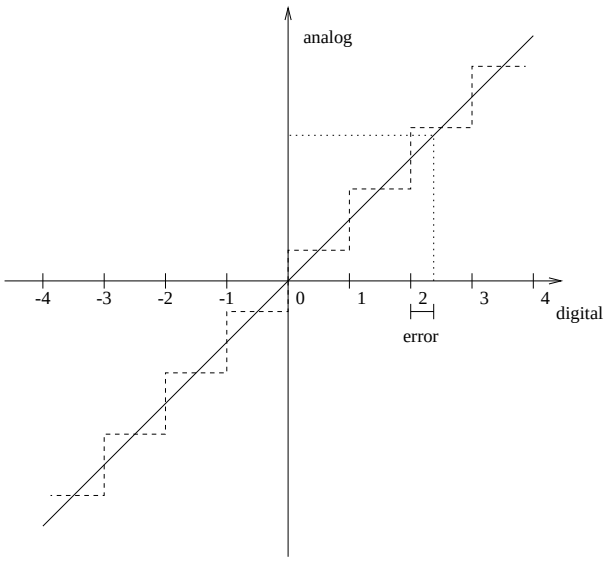


Figure 5.1: Quantisation stair

The root mean square of such a rectangular distributed signal is:

$$x_{rms} = \sqrt{\int_{-\Delta_{max}}^{\Delta_{max}} t^2 dt} = \sqrt{\frac{\Delta_{max}^3}{3}} = \frac{LSB}{\sqrt{12}}\,[V]$$

The power of this noise is randomly distributed over frequencies, thus white noise. The power spectral density of the quantisation noise is:

$$PSD_{quantisation\ noise} = \sqrt{\frac{0.244}{3 \times 40e6}} = 22.29\,[nV/\sqrt{Hz}]$$

This is distributed over all frequencies positive as negative, some people prefers only to think of positive frequencies, for them the PSD would be $31.5[nV/\sqrt{Hz}]$

This noise is equally distributed over the all of the frequency range, that is from negative Nyquist frequency to the positive Nyquist frequency. If we had a sine wave, with an amplitude that just saturates the ADC, then we would have a signal-to-noise ratio of:

$$SNR = 20log\frac{1/\sqrt{2}}{LSB/\sqrt{12}} = 20log\frac{1/\sqrt{2}}{2/(2^N-1) \times \sqrt{12}} \approx 6.02N + 1.76$$

NOTE that the approximation is only because of the ignoring of the -1 in the dominator. In the case of high resolution this constant is insignificant (i.e $2^{12} = 4096 \gg 1$).

This noise is white and thus distributed over all frequencies. In our case we only look on the band from 5.5 to 6.5$[MHZ]$. Thus a bandwidth, BW, of only $1\,[MHz]$. The noise outside this band, do not influence our calculations. By zooming in on a more narrow band, we get a SNR of:

$$SNR = 6.02N + 1.76 + 10log\frac{f_s}{BW}$$

The quantisation noise with our specifications corresponds to an ADC with the following number of bits for representation.

$$\tilde{N} = N + \frac{10log\frac{40e6}{1e6}}{6.02} = 14.66$$

5.1.1 Spurious

The 'stair-like' quantisation on fig 5.1, is an ideal situation. In the real world, there is not equidistant distribution of digital levels for all frequencies. This is, among others, due to unideal sample and hold of the incoming analog signal. The effect is distortion of the frequency spectrum.

As we're making a digital zoom on the frequencies of interest, we could in theory forget all other frequency components outside this band. But due to these spurious frequency components, we might have components that interfere with our band of interest. As this band consist of weak contributions, these artifacts could endanger our Schottky information. The spurious components are from the specifications given to less than $-80[dB]$ of the signal. This is the Analog Devices $80\,dB$ spurious Free Dynamic Range (SFDR). If the spurious components disturbs the baseband then a, dithering technique can be used. The technique basically just consist of adding low band noise to the signal. This noise is out of baseband and doesn't disturb the interesting frequencies. Spurious components are however smeared or dithered and becomes part of the noise floor, with a significantly smaller variance. This system probably needs such a technique to improve the baseband, but no profound study of it is made, yet. Further information about it, can be found in [9].

5.2 Mixing principle

Real signals

A discrete real sinusoidal oscillations has the frequency spectrum:

$$g(n) = asin(2\pi f_1 n\Delta T + \phi)$$

$$g(n) \leftrightarrow \begin{cases} G(f) = (a/2)exp\{\pm j\phi\} & f = p(f_s \pm f_1) \quad n, p \in N \\ G(f) = 0 & elsewhere \end{cases}$$

Table 5.1: Frequency components from analog downmixing

f	$G_{mix(f)}$
$f_1 - f_0$	$(a/2)exp(j(\phi + \frac{\pi}{2}))$
$-f_1 + f_0$	$(a/2)exp(-j(\phi + \frac{\pi}{2}))$
$f_1 + f_0$	$(a/2)exp(j(\phi + \frac{\pi}{2}))$
$-f_1 - f_0$	$(a/2)exp(-j(\phi + \frac{\pi}{2}))$

, where f_s denotes the sampling frequency and $\leftrightarrow$ the Fourier transformation with time interval T . Mixing it with another real sinusoidal function creates a signal, which has a frequency spectrum consisting of the convolution between the two.

$$\begin{aligned} g_{mix}(n) &= asin(2\pi f_1 n\Delta T + \phi)sin(2\pi f_0 n\Delta T) \\ &= (a/2)(cos(2\pi(f1 - f0)n\Delta T + \phi) \\ &\quad -cos(2\pi(f1 + f0)n\Delta T + \phi)) \qquad [Schaum\,5.65] \\ &= (a/2)(sin(2\pi(f1 - f0)n\Delta T + \phi + \frac{\pi}{2}) \\ &\quad -sin(2\pi(f1 + f0)n\Delta T + \phi + \frac{\pi}{2})) \end{aligned}$$

This signal, $g_{mix}(t)$, has the frequency components written in table 5.1

This corresponds to the fact that real signals, always has the relation between negative and positive frequencies of:

$$G^*(f) = G(-f)$$

,where $G^*(f)$ denotes the complex conjugate of $G(f)$.

This means that if you want to move a frequency of, say $10[Hz]$ down to $5[Hz]$, then you can multiply with a sine of $5[Hz]$. Thus by convolution you get your wanted $5[Hz]$ component, but due to positive frequencies you get a component at $15[Hz]$, as well. If you want to avoid this component from the positive frequency, then complex signals must be introduced.

Complex signals

For a complex signal, the frequency spectrum after mixing is slightly different. A mixing in a complex mode is done by multiplying with a complex frequency.

$$g_{cmix}(n) = g(n)exp\{j2\pi f_0 n\Delta T\} = g(n)cos(2\pi f_0 n\Delta T) + jg(t)sin(2\pi f_0 n\Delta T)$$

Table 5.2: Frequency components from complex downmixing

f	calculations	$G_{cmix}(f)$
$f_1 - f_0$	$(a/2)(exp(j\phi) - exp(j\phi)) =$	0
$-f_1 + f_0$	$(a/2)(exp(-j\phi) - exp(-j\phi + \pi)) =$	$aexp(-j\phi)$
$f_1 + f_0$	$(a/2)(exp(j\phi) - exp(j\phi + \pi)) =$	$aexp(j\phi)$
$-f_1 - f_0$	$(a/2)(exp(-j\phi) - exp(-j\phi)) =$	0

using the relation between real and complex frequencies:

$$asin(\theta) = \frac{a(expj\theta - exp{-}j\theta)}{j2} \qquad [Schaum\,7.17]$$

we can decompose the complex mixed signal into two signals. The real and imaginary part of only real sine frequencies. From the rule of linearity[1] we can solve them isolated.

$$\begin{aligned}
g_{cmix}(n) &= \frac{a}{j2}(exp(j2\pi(f_1 - f_0)n\Delta T + \phi) - exp(-j2\pi(f_1 + f_0)n\Delta T - \phi)f) \\
&= \frac{a}{j2}(cos(2\pi(f_1 - f_0)n\Delta T + \phi) + jsin(2\pi(f_1 - f_0)n\Delta T + \phi) \\
&\quad -cos(-2\pi(f_1 + f_0)n\Delta T - \phi) - jsin(-2\pi(f_1 + f_0)n\Delta T - \phi)) \\
&= \frac{a}{j2}(sin(2\pi(f_1 - f_0)n\Delta T + \phi - \frac{\pi}{2}) + jsin(2\pi(f_1 - f_0)n\Delta T + \phi) \\
&\quad -sin(-2\pi(f_1 + f_0)n\Delta T - \phi - \frac{\pi}{2}) - jsin(-2\pi(f_1 + f_0)n\Delta T - \phi)) \\
&= \frac{a}{j2}(sin(2\pi(f_1 - f_0)n\Delta T + \phi - \frac{\pi}{2}) + jsin(2\pi(f_1 - f_0)n\Delta T + \phi) \\
&\quad +sin(2\pi(f_1 + f_0)n\Delta T + \phi + \frac{\pi}{2}) + jsin(2\pi(f_1 + f_0)n\Delta T + \phi)) \\
&= \frac{a}{2}(sin(2\pi(f_1 - f_0)n\Delta T + \phi) + sin(2\pi(f_1 + f_0)n\Delta T + \phi) \\
&\quad -jsin(2\pi(f_1 + f_0)n\Delta T + \phi + \frac{\pi}{2}) \\
&\quad -jsin(2\pi(f_1 - f_0)n\Delta T + \phi - \frac{\pi}{2})) \\
&= \frac{a}{2}(g_1(n) + g_2(n) + jg_3(n) + jg_4(n)) \\
&\leftrightarrow \quad G_c(f) = \frac{a}{2}(G_1(f) + G_2(f) + jG_3(f) + jG_4(f))
\end{aligned}$$

This is four contributions and transformed isolated and summed in frequency in table 5.2:

[1] $ag_1(n) + bg_2(n) \leftrightarrow aG_1(f) + bG_2(f)$

As shown, we get only spectral components convoluted with the positive image frequency.

A comparison of a Fourier transformation of these two techniques are shown on figure 5.2.

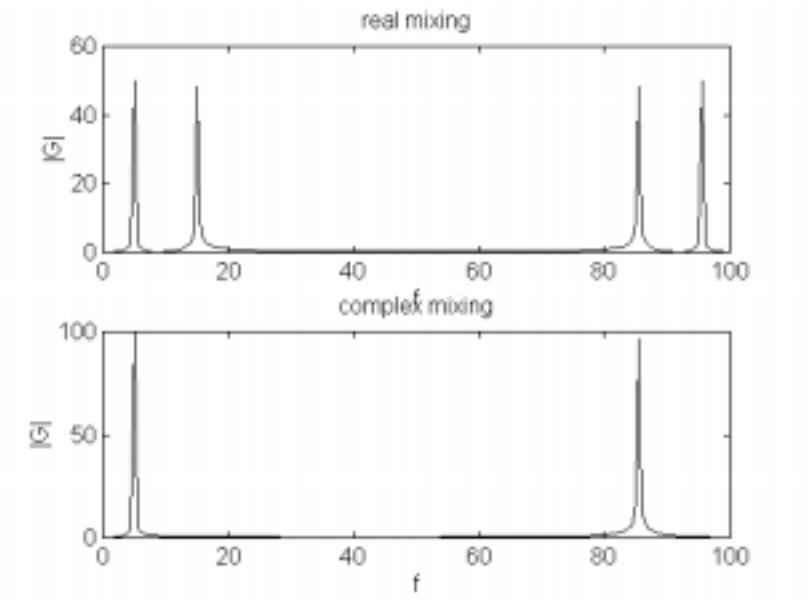


Figure 5.2: Spectra from real/complex mixing

5.2.1 Downmixing

A very important feature of this technique, is that the negative image frequencies are completely unimportant for the mixing. If a mixing with a real frequency where done, then a band would distort when downmixed, because of these negative image frequencies. In a complex mode, though, the frequency spectrum is moved according to the formula:

$$g(n)exp(-j2\pi f_0 n\Delta T) \leftrightarrow G(f + f_0)$$

So having a baseband, around a centre frequency, lying at a frequency compared to the baseband, it is profitable to downmix this baseband to lower frequencies, thus decreasing the number of samples.

The downmixing of N samples takes N arithmetical operations with the complex frequency, thus $2N$ real arithmetical operations. Then the signal has to be lowpass filtered with say p coefficients, thus introducing additionally $2Np$ arithmetical operations. The signal doesn't need this many samples to be decoded, now. As the interesting signal is limited by

width of the baseband we now dont care about frequencies above, say[2] f_g. So we only need samples with a time distance of $\frac{1}{f_g}$. The factor reduced is thus $m = \frac{f_s}{f_g}$. Further more we do not need to low pass filter more than these samples, thus reducing the low pass filtering to $\frac{2Np}{m}$ arithmetical operations. The total number of arithmetical operations is now less than $Nlog_2N$ for large values of m, namely:

$$2N + \frac{2Np}{m} + \frac{N}{m}log_2\frac{N}{m}$$

Normal FFT algorithms requires $Nlog_2N$ arithmetical operations, so just doing one 256-point FFT with a sample rate of 20 [MHz] would require a clock frequency of $160[MHz]$ if no parallel calculations were possible[3]. So this is a big calculation load and weare really only interested in knowing a certain baseband, around the revolution frequency or a harmonic of this. So even with only a DSP it would be profitable to downconvert our signal to baseband before Fourier transforming the input signal. In our system it is much more efficient because of the fact that the two tasks are performed by two different arithmetic units in parallel.

5.3 FFT

The Fourier transform of a discrete time signal can be done in several ways. A straightforward way is by performing a **Discrete Fourier Transform, DFT**:

$$X(k) = \sum_{n=0}^{N} x(n)W_N^{nk}$$

,where W_N^{nk} is the twiddle factor given as $W_N = exp\{\frac{-j2\pi}{N}\}$. This way of calculating the Fourier transform takes N^2 complex multiplications and complex additions. There is a way to reorganise the multiplications and additions to obtain a faster calculation and this is done by the **Fast Fourier Transform, FFT**, algorithm.

NOTE that as the twiddle factor is cyclic the algorithm automatically presumes that the same signal section is repeated infinitely. When a cyclic nature of a signal has an odd ratio with the period analysed, then the FFT introduces frequencies that are not present in the signal. This effect is less significant when a window function is used, see section 3.5.

[2]baseband originally being between $f_0 \pm f_g$

[3]In DSP parallel arithmetic is possible, TMS320C40 carry out max. 2 per clock

DIT/DIF

The principle of the FFT algorithm, is to divide the task into smaller ones and to use some intermediate results common for several frequency components. The FFT originates from the DFT algorithm, but uses such intermediate variables. These variables can be defined either beginning from the time domain or frequency domain. The two principles are respectively called **Decimating In Time, DIT, and Decimating In Frequency, DIF**. The number of calculations is exactly the same for both algorithms, it is only a matter of the order of calculations in the algorithm.

Radix

When dividing these tasks into smaller ones, we end up with a calculation unit which is identical for all calculation layers. This unit is called a butterfly and the size of this butterfly is called the radix of the algorithm. So an algorithm dividing the total DFT calculation task down to recursive tasks of 2- input butterflies, is a radix-2 algorithm. A butterfly is actually a full FFT, but made of a small number of inputs, normally 2,4 or 8 (most frequently 2)

Complex/Real

Further more the input signal can consist of either pure real signals or being complex. But according to the Fourier rule of linearity, this doesn't affect the algorithm:

$$\begin{aligned} Re[x(t)] + jIm[x(t)] &= a(t) + jb(t) \leftrightarrow A(f) + jB(f) \\ A(f) + jB(f) &= FFT(Re[x(t)]) + jFFT(Im[x(t)]) \end{aligned}$$

It can still be performed by making two separate FFTs of respectively the real and complex part of $x(t)$.

If we split up the input signal in two, a real and complex part, then the relation to original complex-signal, $x(t)$ and its transformation, $X(f)$ is

$$Re[X(f)] = Re[FFT(Re[x(t)])] - Im[FFT(Im[x(t)])]$$
$$Im[X(f)] = Im[FFT(Re[x(t)])] + Re[FFT(Im[x(t)])]$$

An FFT algorithm is completely described by the tree options [real/complex], [DIT/DIF] and [radix]. Of course, it can be implement this in a lot of ways, but the algorithm structure stays the same.

5.3.1 Splitting-up into butterflies

DIT

When modifying the DFT algorithm to a DIT FFT we start dividing the discrete time vector into even and odd time indexed signals:

$$x_k = [x_0\ x_1\ x_2\ x_3\ ...\ x_N] = x_{even} + x_{odd} = [x_0\ x_2\ ...\ x_{N-1}] + [x1\ x3\ ...\ xN]$$

The DFT can thus be transformed into:

$$\begin{aligned} X(k) &= \sum_{n=0}^{N} x[n]W_N^{nk} \\ &= \sum_{n=0}^{N/2-1} x[2n]W_N^{2nk} + \sum_{n=0}^{N/2-1} x[2n+1]W_N^{(2n+1)k} \\ &= \sum_{n=0}^{N/2-1} x[2n]W_N^{2nk} + W_N^k \sum_{n=0}^{N/2-1} x[2n+1]W_N^{2nk} \end{aligned}$$

with $W_N = exp\{-j2\pi/N\}$ and $W_N^2 = exp\{-j2\pi 2/N\}$ we see that $W_N^2 = W_{N/2}$

$$X(k) = \sum_{n=0}^{N/2-1} x[2n]W_{N/2}^{nk} + W_N^k \sum_{n=0}^{N/2-1} x[2n+1]W_{N/2}^{nk}$$

This is a two N/2 - point FFTs of respectively the even and odd part of $x(n)$, the odd result multiplied with a twiddle factor. Apart from the two N/2 - point FFTs, we need to multiply the odd FFT with a complex twiddle factor thus introducing N complex multiplications. The combinations of two N/2 point FFTs to a total N- point FFT require N complex additions. So in all, including the FFT calculations we need $N + 2(N/2)^2 = N + N^2/2$ calculations as opposed to N^2 for a clean N-point straightforward DFT. This technique is thus already profitable from $N > 2$. Further more we can again recursive divide the N/2 - point FFT into two other N/4 - point FFTs and continue dividing until the smallest wanted butterfly size is reached. Then there will be $log_{radix}(N)$ divisions of FFT lengths and the total amount of both complex multiplications and complex additions, will be $Nlog_{radix}(N)$.

In the case that the size of input is smaller than N then the x vector must be extended with zeros.

DIF

When decimating in frequency, we start with dividing the frequency components instead. Taking even and odd indexed frequency components the

algorithms look as.

$$\begin{aligned} X[2r] &= \sum_{n=0}^{N-1} x[n]W_N^{n2r} \\ &= \sum_{n=0}^{N/2-1} x[n]W_N^{n2r} + \sum_{n=N/2}^{N-1} x[n]W_N^{n2r} \\ &= \sum_{n=0}^{N/2-1} x[n]W_N^{n2r} + \sum_{n=0}^{N/2-1} x[n+N/2]W_N^{2r(n+N/2)} \\ &= \sum_{n=0}^{N/2-1} (x[n] + x[n+N/2])W_{N/2}^{rn} \end{aligned}$$

The odd part is calculated the same way to

$$X[2r+1] = \sum_{n=0}^{N/2-1} (x[n] - x[n+N/2])W_N^n W_{N/2}^{rn}$$

These two frequency equations are DFTs of respectively a signal $a = x[n]+x[n+N/2]$ and $b = x[n]-x[n+N/2]$, where b is multiplied with a frequency dependent twiddle factor. So as before the division is continued until the smallest calculation unit is reached, again resulting in $Nlog_{radix}(N)$ complex multiplications and complex additions.

Description of an 8-point complex DIF radix-2 FFT

Each butterfly is identical and as it is a radix-2 algorithm, there is two complex inputs and two complex outputs. The algorithm is as written in the DIF section above.

$$X_{even} = (x_1 + x_2)X_{odd} = (x_1 - x_2)W^k$$

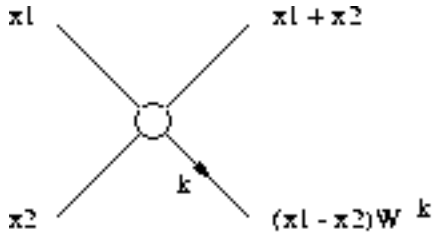


Figure 5.3: DIF butterfly

Knowing that it is an 8-point FFT we need two $(log_2(8)-1)$ intermediate variables to calculate the Fourier transform.

$$INPUT \Rightarrow INT1 \Rightarrow INT2 \Rightarrow FOURIER$$

($\Rightarrow$ signifying N/2 butterflies)

So if the input vector is chronologically ordered as:

$$INPUT = [x_0\ x_1\ x_2\ x_3\ x_4\ x_5\ ...\ x_N]$$

then the first butterfly would look like:

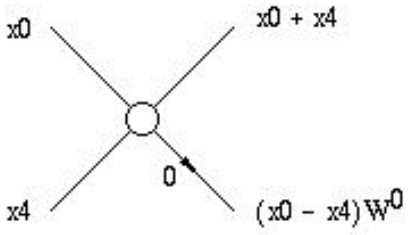


Figure 5.4: first butterfly

The final combination of all butterflies with their twiddle factor exponents is shown on figure 5.5.

5.3.2 Implementation of FFT

When implementing such a structure it is important to profit from the fact that all butterflies are alike and thus can be implemented recursive. Another important factor is the use of memory, it takes a while to get and put data from memory compared to the arithmetic operations. In order to minimise use of memory, the input variables to a butterfly is stored on the same location, as there is not any further use of it. The twiddle factors are stored in memory, as a $1\frac{1}{4}$ sine wave. The reason for this $\frac{1}{4}$ is that there is a need of one period of a sine and one of a cosine. By giving the sine an offset, we obtain a cosine and do not need to change the calculation structure. Basically we only need $\frac{1}{4}$ of a sine, the rest is redundant, but as memory availability is not a problem we avoid calculation by storing it all.

Complex butterfly implementation

A complex input vector is ordered as an array with successive real and imaginary parts, starting with the real. So a complex butterfly will have four real inputs for a radix-2 algorithm. As only real operations are possible, in low level programming, we must decompose the compact form shown in the butterfly on figure 5.6.

x1 and x2 is here complex representative of respectively

$$x1 = x_{Real} + jx1_{Imag}$$

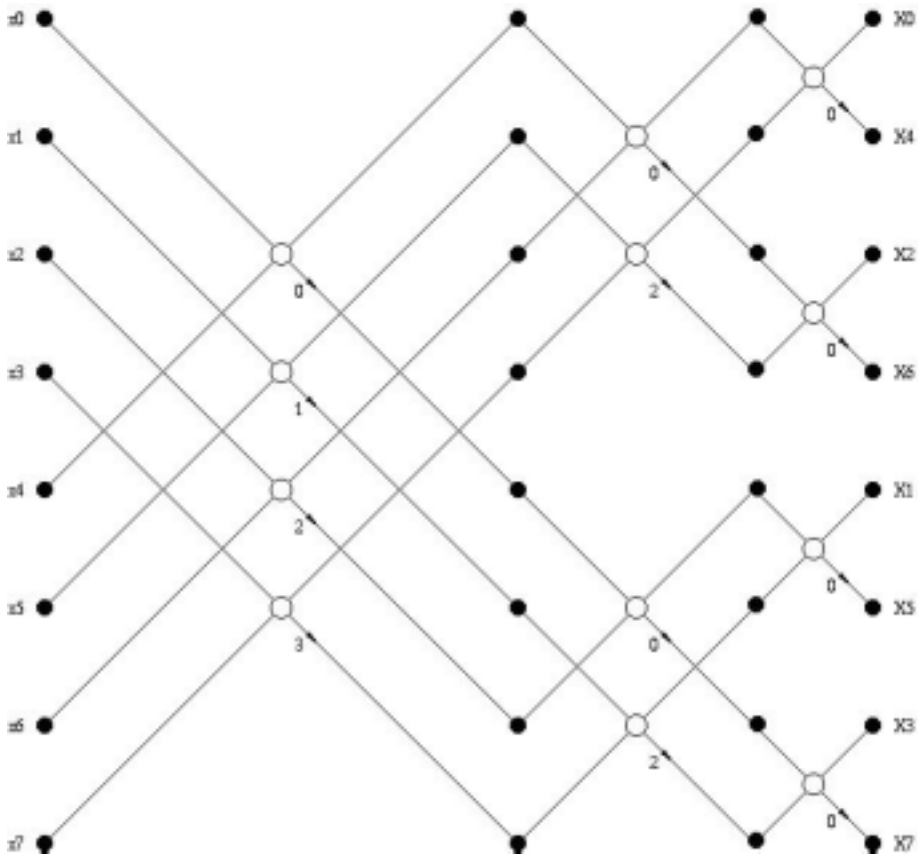


Figure 5.5: 8-point DIF FFT

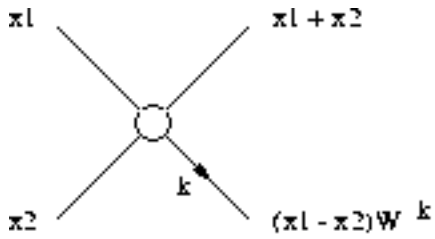


Figure 5.6: complex butterfly

$$x2 = x2_{Real} + jx2_{Imag}$$

The result has thus complex values, which is divided into to cells a real and an imaginary one.

$$\begin{aligned} \mathbf{x1 + x2} &= Re[x1 + x2] + jIm[x1 + x2] \\ Re[x1 + x2] &= x1_{Real} + x2_{Real} \\ Im[x1 + x2] &= x1_{Imag} + x2_{Imag} \end{aligned}$$

$$
\begin{aligned}
(\mathbf{x1} - \mathbf{x2})\mathbf{W^k} &= Re[(x1 - x2)W^k] + jIm[(x1 - x2)W^k] \\
Re[(x1 - x2)W^k] &= Re[(x1 - x2)cos(k) + j(x1 - x2)sin(k)] \\
&= (x1_{Real} - x2_{Real})cos(k) - (x1_{Imag} - x2_{Imag})sin(k) \\
Im[(x1 - x2)W^k] &= Im[(x1 - x2)cos(k) + j(x1 - x2)sin(k)] \\
&= (x1_{Imag} - x2_{Imag})cos(k) + (x1_{Real} - x2_{Real})sin(k)
\end{aligned}
$$

(NOTE that the TI FFT algorithm has different signs of the sine function related to the definition of k in the butterfly, $sin(-k) = -sin(k)$, $cos(-k) = cos(k)$)

The call to a complex butterfly function has 3 pointers and one displacement value for the COSINE generation. Two of the pointers are pointing at the input values and the third at the SINE table. The two input pointers are pointing at the real part of the input, the subsequent memory location contains the imaginary part, see figure 5.7.

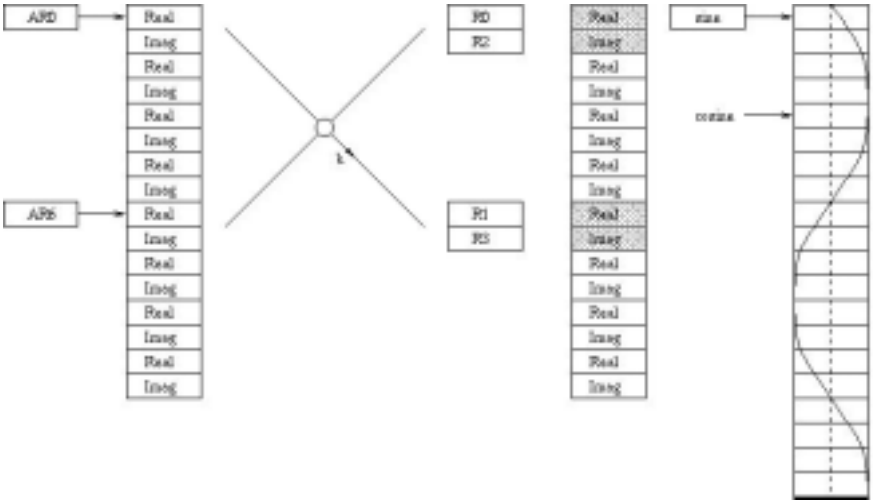


Figure 5.7: Use of memory cells for FFT calculation

For intermediate values we use 6 registers, the R0, R1, R2, R3,R4,R5 extended registers. The input is the two complex signals, upper input UR+jUI and lower input LR+jLI. These inputs are overwritten, by the results of the butterfly. They can NOT be overwritten, before the use of them is fully finished. As we do not want to use too many registers, we need a certain timing of instructions, to avoid overwriting wrong values. The approach taken in a TI FFT algorithm is visualised in the figure 5.8.

The timing is divided into phases, symbolised by vertical punctured lines. Everything happening in a phase can happen in the order the programmer chose. If however, the former phase is not completely finished,

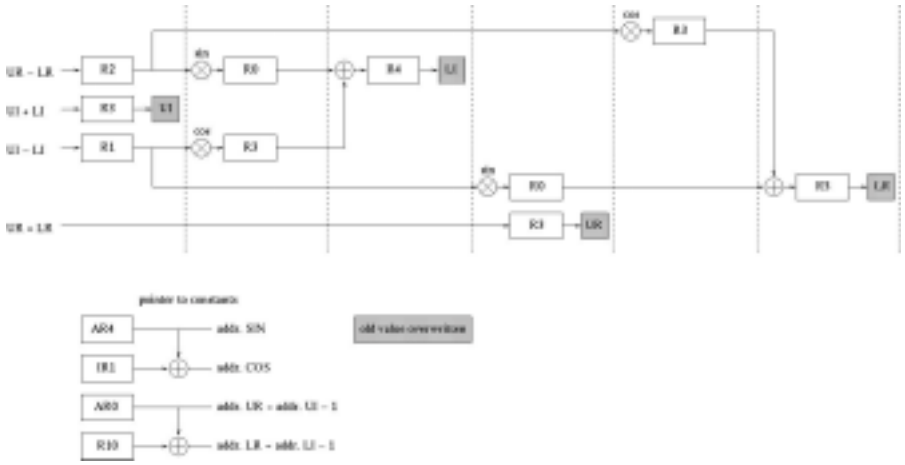


Figure 5.8: Use of registers for butterfly

one should be cautious not to overwrite wrong values.

Registers used for addressing of input values and twiddle factors are shown at the bottom of the figure. The two complex inputs have the imaginary part subsequent to the real.

A straight forward implementation follows:

∗AR0	=	PTR(UR)
∗AR6	=	PTR(LR)
∗AR4	=	PTR(SINE)
∗IR1	=	displacement SINE vs. COSINE ⇒
PTR(COSINE)	=	*+AR4(IR1)
SUBF	*AR6,*AR0,R2	;UR - LR → R2
ADDF	*+AR0,*+AR6,R3	;UI + LI → R3
SUBF	*+AR6,*+AR0,R1	;UI - LI → R1
STF	R3,*+AR0	;R3 → UI
MPYF	R2,*AR4,R0	;R0 := R2 SIN
MPYF	R1,*+AR4(IR1),R3	;R3 := R1 COS
SUBF	R0,R3,R4	;R4 := R3 - R0
STF	R4,*+AR6	;R4 → LI
MPYF	R1,*AR4,R0	;R0 := R1 SIN
ADDF	*AR0,*AR6,R3	;R3 := UR + LR
STF	R3,*AR0	;R3 → UR
MPYF	R2,*+AR4(IR1),R3	;R3 := R2 COS
ADDF	R3,R0,R5	;R5 := R3 + R0
STF	R5,*+AR0	;R5 → UI

This is the basic butterfly, with input parameters passed in AR0,AR6,AR4 and IR1. This one takes 14 instruction, which are all needed. However the

DSP structure has a several CPUs, each taking care of their field (arithmetic and addressing). So some instructions can take place in parallel. This can be reduced to 9 instructions, only changing the order and putting in parallel instructions. As this butterfly loop is repeated 769 times[4] for a 256 point FFT, these 9 instruction occupy the majority of the assembly function. By rewriting only 14 lines the speed improves around 35%.

This complex butterfly is either called or integrated in a loop. The loop is changing index of input and the argument of the sine function as well as keeping track of what level is being calculated. Thus, the first 3 calling parameters (AR0, AR6, AR4). The principle of the algorithm changing these parameters, is coded in the algorithm beneath.

The TI algorithm use the fact that quite a lot of butterflies has a zero twiddle factor. In principle the same butterfly can be used, but it would result in useless multiplications with 1 and zero. In stead they have made a zero butterfly and a non zero one. They are then integrated in a loop structure that increments twiddle factors, input pointers etc.

The Texas Instruments FFT algorithm is shown in appendix M with corrections made by the author of this report. Every correction has a !! symbol in the beginning of the comment.

5.3.3 Storing FFT results

The FFT calculations are carried out in 40 bit registers, but data are quantified in 12 bit. The following section enables a comparison of resolution between a high resolution FFT and the FFT done by the 'C40.

When a number is quantified, there is a loss of precision. A quantisation is normally modelled as the right number plus a quantisation error, $\hat{x} = x+\epsilon$. Presuming that the number is quantified by rounding towards nearest value the maximum error is:

$$\epsilon = \frac{2\,[V]}{2 * 2^{12}} \approx \pm 0.24\,[mV]$$

This error is completely random, thus a white noise contribution.

The resolution in the DSP is a lot higher than this ADC. When storing data, the resolution is stored in 32 bits. Internal calculations can be performed in 40 bit registers, such that 32 bits of resolution is preserved when performing arithmetic on data.

[4]The butterfly with 0 twiddle factor is calculated differently, so 128-1+128-2+128-4...128-128 = 769 calls with non-zero twiddle factor

Table 5.3: 2's complement cut-off example

2's complement	in decimal	when LSB cut-off	error
011	3	2	-1
010	2	2	0
001	1	0	-1
000	0	0	0
111	-1	-2	-1
110	-2	-2	0
101	-3	-4	-1
100	-4	-4	0

In the 32 bit registers we store numbers in floating point format. This means that we reserve 8 bits for the exponent and 24 for the mantissa. The mantissa, however, including the sign of the number. In TI floating point representation this leaves us 23 bits for representing a fraction between 0.5 and 1.

For a positive number there is 23 bits at our disposal to quantify 0 to 1 Volt. When storing from extended precision registers to 32 bit registers the last bits is just cut off. Introduces a negative error between 0 and $-\frac{1}{8}$ ppm.

For negative numbers the error introduced is similar, as the fraction is represented in 2's complement numbers. See example in table 5.3.

The absolute error of the 32-bit floating point number is:

$$\epsilon_r =< \frac{\epsilon}{fraction} >= \frac{< \epsilon >}{< fraction >} = \frac{-1/8}{3/4}\ ppm. = -\frac{1}{6}\ ppm$$

So every time we store a number represented in a higher resolution than 32 bits, we introduce an error of $-\frac{1}{6}\ ppm$.

These calculations correspond to a check of the assembly FFT algorithm with the sim4x TI 'C40 simulator. The algorithm was checked up against the Matlab 15 decimal digit resolution (64-bit floating point) and the difference in spectrum corresponded to the calculations above. Compared to the C40 resolution the Matlab results can easily represent the true value, denoted x above. The errors will not be accumulated by multiplications or additions they only occur when the data is stored in 32-bit memory.

This means that if we take 1000 numbers from the 32-bit memory, multiply them with some factor, then we loose resolution by storing them in 40-bit registers. But this is a random error. When we put it back

into 32-bit memory we throw away the last bits and adds a biased error of $-\frac{1}{6}$ *ppm* in mean. Performing an accumulation in extended precision registers will at the end give a biased error of $-\frac{1}{60}$ % which is no longer in LSB. This does not threaten our resolution, but one should be aware of this tiny biased error when comparing results.

Chapter 6

Development

This chapter covers, in great lines, the procedure which this project has followed. The system has been build up from scratch. No previous experience in such types of acquisition and processing systems was present at CERN before.

The chapter starts out with covering the different phases of the system development. This involves description of information gathering, hardware decisions, development approach etc. All of the phases important to avoid unsuccessful project outcome. The aid of specific tools, is essential. A section covers the most important of, those used for this project. A small section goes through the purchasing of project material. This is basically because it has been an important parameter, for the course of this project.

6.1 Development procedure

The development, of the data acquisition and processing system, is done in many phases. Parallel to the data acquisition and processing system development, the rest of the AD project system is developed. Most of the development is done isolated from the final environment.

The development phases are written in this section in a chronological order.

Choosing equipment

The first thing done was gathering of information from different companies producing hardware suitable for our purposes. This was mainly done by my supervisor, Flemming Pedersen before my arrival at CERN. There where really only two companies that produced digital receivers for crates and Pentek was the most reliable on the market. Further more the other company didn't have as big a variety of modules as Pentek had. CERN has a long history of projects relying on unfortunate products. Company

shut-down, bad support, software and hardware bugs etc. are some of the obstacles, which should be avoided. An older beam diagnotics system, was f.x. based on a Motoriola DSP which is now basically off the market, thus no more support. The Texas Instruments DSP seems to have a better market share, than Motorola had. The choice was done on the basis of these facts.

The models from Pentek was, however, not chosen before my contract started at CERN. Through meetings with vendors and studies of model specifications, the models where picked and ordered.

In parallel, the search of software was carried out mainly through the Internet. Enquiries about compatibility and hardware requirements was made, to several companies, by email and fax. Their products were studied from information available on the WEB. Through advises from vendors, distributors and people working in these environments we chose the software constellation and asked for commercial offers from different distributors

Acquiring knowledge

To facilitate the acquiring of knowledge me and my supervisor took a 5 day Texas Instruments course, at Cranfield University(UK) in the TMS320C40 processor. This and studies on my own of manuals and course material consisted of the acquired knowledge of the processor.

The rest of the system was studied by inspection of manuals and tutorials from commercial material.

Simulation

With the simulator, which arrived 2 months before the hardware, the first code was run. The compiler and linker arrived at the same time, so in an artificial environment, the first pieces of code was be tried out. A small version of the complete processing software was build in assembly and C and run on the simulator. The simulator is, however, thousands of times slower, so only program principles could be checked this way. Larger statistical analysis was not possible in this environment.

First hardware tests

The hardware was at first tested in the PS/RF laboratory, room 561-R01. We arranged a VME crate and a PC for debugging purposes. The GO-DSP Code Composer was used to compile, assemble, link and debug. Some features of the Code Composer enables a far more thorough inspection of the code, than the simple features of the Texas Instruments simulator, sim4x.

Implemented hardware tests

Later the hardware is going to be installed in the final crates, in the control room of the AD, building 193. From here, the modules are still connected via a ∼1 meter long J-TAG cable. Physically the environment is still as it were in the laboratory, only with other power supplies.

Test on beam

When the pick-ups are in place and a beam is available, then the real signals can be connected to the ADC module input. The pick-ups are scheduled to be in place around the 1st of November. There is no general timing available, as it is still an isolated environment with only one input, the beam signal. Some of the external interrupts needed for the processing is simulated manually.

When the system is tested and has shown the required performance, the last parts of the modules and spares are ordered.

System implemented in network

The embedded data acquisition and processing system is implemented in the control system network by software written for the DSC. This software is scheduled to be written in January and February 1999, by CERN staff employee Georges-Henry Hemelsoet.

System controlled by workstations

When there is access from the network to the modules, user interface software can be written for UNIX workstations. The person responsible for this task is still not found.

Expansion of the system

A possible expansion of the system is carried out, if there is a need of more processing power. DSP boards from Pentek are already studied, so the option of expansion is always held open. For such a new acquisition there will only be a minimum of development effort involved. The code is already prepared for this task, as the only thing, in principle, is to change the settings of the RTOS which are already used for the single processor system. There is a need of expanding the 3L Diamond RTOS for multiprocessor systems, but this does not change the work effort.

6.2 Development tools

In this section the used software tools are introduced. The available computer hardware was in the office a UNIX workstation and a Pentium PC running Windows 95. At the laboratory a 486 PC with the same operating system. The workstation is connected to the CERNSP computer network and the PC to the CERN NICE network. In the following the software tools, running on these machines, will be introduced. There is a brief introduction to the use of the less known tools, but the development use is described for every one of them.

6.2.1 Matlab

As tool of mathematics, Matlab was used. It was used to verify theoretical formulas, analyse signal processing principles and for minor or larger system simulation. An important role was the import/export of values between the Code Composer environment and Matlab. Around 50 smaller or larger m-files has been build during the project for studies of a variety of problems. This tool has been a helpful companionship.

The outcome for the reader of the use of this tool, is verifications of different equations and text supported by graphs. None of the Matlab m-files are enclosed in the appendices as there will be no further use of them.

6.2.2 Sim4x

The simulator was used to develop and verify code in an office environment. It came before the real hardware, so in the beginning it was the only mean of testing. It was, however, the intention to have an option of developing new code, in a calm environment, in stead of the less private laboratory or the noisy control room.

The simulator for the TMS320C40 and TMS320C44 is a simple old simulator called Sim4x. It is an older product from 1993 or before and it was even shipped on an old 5 1/4 inch disc. The screen environment is of the old DOS format, so a low text resolution limits the amount of information on the DOS screen. The program is based on commands from the prompt, but a lot of things can be carried out by aid of the mouse. It is an isolated environment that only enables data analysis from the screen and not export of data, to other more power full analysis tools. It is, however, possible to COPY and PASTE via a MS-DOS shell window, but this is only recommended to do, with small data series (it is very time consuming). The best way to copy data, this way, is by maximising the memory window, in floating point format, thus showing the maximum 88 values. So the simulator is

mainly used for qualitative analysis and to limited extend to quantitative analysis. Some data is copied to text files via this copy and paste procedure, this enabled analysis of data carried out in a Matlab environment.

The program is however quite simple to users familiar with older software types. There is only a limited variety of possible options. So very fast any frequent user will have visited almost any corner of the simulator. Because of the fact that a simulation of a 20 msecs DSP process, takes close to 2 minutes, you get a natural feeling of where the time consuming code is.

Screen environment

The visible area on the DOS screen is only the usual 80×25 characters. This limits the amount of windows visible at the same time. By default, there is the COMMAND, MEMORY, DISASSEMBLY and CPU window shown, but a lot of others are possible. The windows are easily modified in size and placement by use of the mouse. However, it is an old interface that do not permit modifications that is usually present in all other newer MS-Windows based software. Some unlucky usage of window displacements, leads to scrambling of the screen and a restart is necessary.

The windows available are described in the subsections following. There is a large number of possible commands that can be written in the command window, for further study of those see [10]. Some of the most important ones will, however, be covered here.

Command window

The most important window is the COMMAND window. From this window all the other window parameters can be controlled. If nothing is done to deliberately make another window text sensitive, one can always write a command even if another window is active (highlighted). The command will be written in a DOS like prompt and when ENTER is hit, the command is performed. Former commands and answers can be found by scrolling upwards in the COMMAND window.

To load a compiled and liked file the *load FILE* command is used. If this is done from the start there is no need of resetting, if however a reset is needed then *reset*, is written at the prompt. In the same way a *restart* can be written if a debugging session needs to be restarted. Another helpful command is *ba ADDRESS*, which puts a breakpoint at the address following the command. The address can be both a label(ex. main) or a 32-bit physical address(ex. 0x002ff800). The list of breakpoints is found by invoking *bl*. To open another memory window surveying a memory section

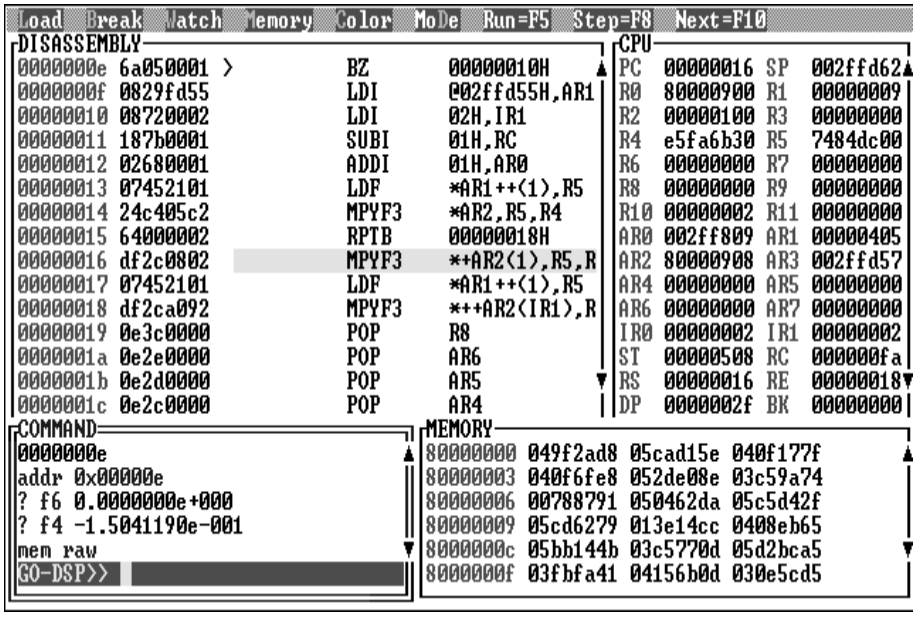


Figure 6.1: Sim4x default screen appearance

in a specific format you invoke f.x. *disp *(float *) ADDRESS* and a window with floating point values will appear. Registers can be surveyed in a separate window, by invoking *wa REGISTER,f* for floating point format representation. An important watch facility is the *wa clk*, which watches the clock cycles performed. Another important usage of watch is to survey the pipeline, this is done by invoking the four commands *wa (void)faddr, wa (void)daddr, wa (void)raddr, wa (void)xaddr.*

As it is often the same set of commands that is repeated, there is an option of writing a log file (extension .log), which is the equivalent to the DOS batch files (extension .bat). This log file is executed by invoking *take LOG-FILE*

Disassembly window

The DISASSEMBLY window can show source code in both C and assembly format. Breakpoints can be set via the left mouse button as well as removed. Scrolling of the command history window, can be done with up down keys[1]. Jumping to a specific address is done by the command *addr ADDRESS*. Debugging step by step is possible by tabbing the key *F8* and jumping to next breakpoint by *F5*. A line can be modified by use of the right button on the mouse, when this is hit the corresponding line appears in a window

[1]scrolling with the mouse is NOT advised as it is too fast

and can be rendered. Illegal modification will leave the line unchanged.

CPU window

The CPU window shows all the registers in the C40. They are all shown in 32-bit format, however the R0-R8 registers are extended precession registers represented by 40 bits. So not all of the information is shown. By invoking *? f0* you get the full floating point representation of register R0. The command *? R0* gives the integer representation. Every registers is highlighted when modified, but if the same value is written to the register it do not highlight. The registers can be changed by double clicking, with the left mouse button on the specific cell making this window text sensitive.

Memory window

The memory is by default shown in hexadecimal format, if the number is floating point, then as the TI floating point format hexadecimal represented. It can be changed to f.x. normal floating point by writing *mem ADDRESS,f*. The values can be changed, the same way as registers, by double clicking the left mouse button and is also left highlighted when modified.

6.2.3 GO-DSP Code Composer

The GO-DSP product, Code Composer, is used for development of the DSP code. This software, interfaces the user with the DSP, in a user friendly way. The compiler, assembler, linker and debugger is integrated in the software and can be controlled by Code Composer. The interface and functionality resembles that of Visual-C. The product is an alternative to the Texas Instruments debugger[2], which is less up-to-date, but has an interface that resembles the simulator, Sim4x. In this section the most basic features of Code Composer is run through. Further information is found in [18].

Graphical User Interface

The graphical user interface is shown on figure 6.2. There are a bunch of windows that can be opened for various tasks. The most important are:

- Project window, showing every source files in the project
- Compiler window, showing errors, the state etc. of compilations
- Memory window, showing blocks of data in optional format
- Dis-assembly window, showing the debugable assembly code

[2]GO-DSP is now owned by TI so one alternative will probably disappear

- C window, showing the debugable C code
- Graphical window, showing plots of memory blocks
- Register window, showing present register values
- Watch window, showing variable or constant values

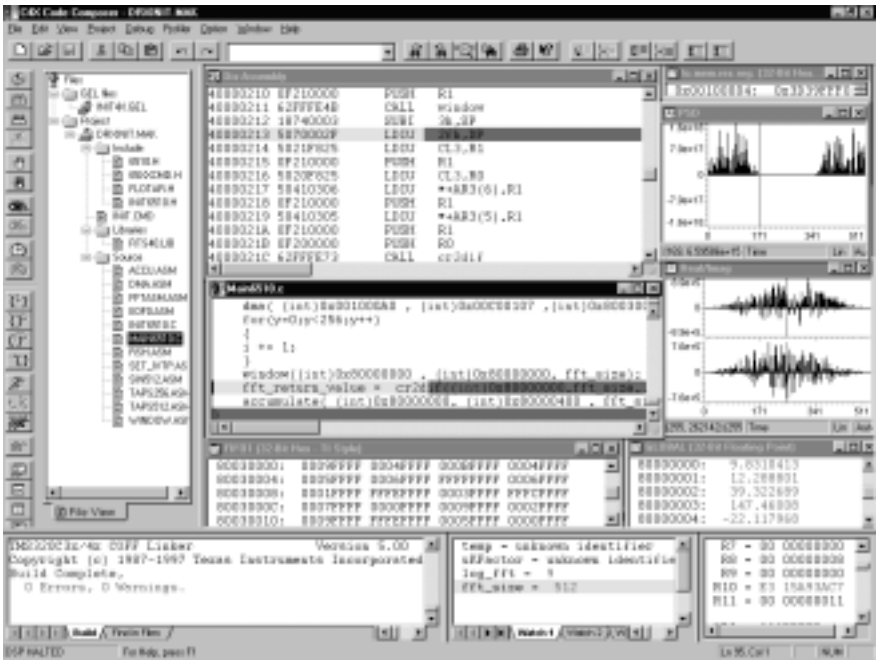


Figure 6.2: Code Composer Graphical User Interface

General for all of the interface, is that memory cells or registers that changes, becomes highlighted red when they are changed. Updating with the same value does not highlight memory. Every memory cell, which is not defined, has the value of the last memory cell that were defined. Numbers showed in the memory window are thus not always existing.

The memory window is editable, so the user has directly on-line access to memory cells and registers. This is an advantage as one does not need to recompile with other register settings every time a new approach is taken. One of the great advantages, is that it is possible to set, for example, an interrupt flag, disable an interrupt, change control registers etc. during the course of a program execution.

The current execution of code is always symbolised by a highlighted yellow bar, when the processor is halted. When halted, new breakpoints etc. can be set or removed. There is a feature called probe-point which is similar to a breakpoint, only it updates the graphical user interface at the

point and continues code execution afterwards. Halting at breakpoint will, however, update the interface as well.

The graphical window has a lot of optional settings, it enables good inspection of larger data series in a fast way. Exact values can be inspected by a sliding data probe. The window is, as well as the others, an object with properties which are quickly changed.

Debugging

The debugger communicates with the TMS320C40 via a J-TAG[3] connection. A J-TAG emulator card, is installed in the development PC. This communication channel is quite slow. Whenever contact to the DSP is required, a time-over lap is needed. Executing one single line (single stepping) of assembly code for example takes about one second or more, where the DSP executing time is only responsible for 25 nano seconds. Whenever a code is released it runs on true clock frequency until the next halting is performed. The halting requires flushing of the pipeline, so time consumption measurements has to be done in a special way, to avoid calculating this flushing overlap. This is however only around 8 clock cycles, so in our case with FFT clock cycle requirements of around 60.000 this can be neglected.

Resetting DSP bug

The setting of the Pentek 6510 DRX board is different from the normal powering up state for the DSP. The local memory is put in a different location, so whenever a reset is performed, the local interface control word register should be changed to:

$$0x00100004: \quad 3D39FFF0$$

It is the "D" in the control word which has to be changed from an "E". We have no mean of changing these power up settings.

32 bit constants bug

Whenever a constant with a most significant bit of one, has to be written it has to be done differently. The program has a bug, that does not enable implementation in hexadecimal. An implementation as

```
value       .word       F000  0000
```

will result in an "INVALID TYPE" error message. However the same constant in binary

```
value       .word       1111  0000  .....
```

[3] Joint Test Action Group

is accepted by the compiler.

Labels

Labels can not be used in memory windows as symbol of addresses. It is necessary to create a map file and open this to find out which is the corresponding label address. Only in the Dis-assembly window, such labels can be used.

TMS3290C40 Assembler,C-Compiler,Linker

The Texas Instruments code development tool for TMS320C4x. These are compiler invoked like all other compilers. They have special settings which are described in [10] and [12]. No further description is found here, for use please refer to the manuals.

6.2.4 3L Diamond RTOS

The software needs to be controlled by a real time operating system (RTOS), for one important reason. If, or when, the system is expanded with another multiple DSP card, see section 4.1.1, then the porting to this new environment is easily done, when using a the same RTOS. A recompilation with new settings is all which is required.

Another advantage is that the structure, of software, is forced to follow a certain common standard. The modules has to be as small as possible and a well defined structure is necessary.

The RTOS has not yet been implemented, in the software version. A small introduction to it is, however, written anyway.

How to use the RTOS

The RTOS can be thought of, as an include directory that organises the calling and interrupt structure. The software modules are still written as they were before, but the communication between modules, are defined in their function calls in another way. This comprises definitions of I/O ports and I/O channels. A configuration file, contains the necessary information of the processor environment. The compiling procedure with a RTOS is the following.

The difference from before is only the last step, with the configurer. The structuring task is done by the RTOS configurer, but some additional information is required. Priorities of each task has to be defined. The task communicate with each other, in a network, that has to be specified in early stages. Then the hardware data and preferences for master DSP etc. has to be defined. With this information, the RTOS configurer distributes the tasks to processors and builds up the calling and interrupt structure.

	compiler		linker		configurer	
∗.c , ∗.asm	→	∗.obj	→	∗.tsk	→	∗.app
source files		object files		task files		application files

Table 6.1: Compiling procedure with 3L Diamond RTOS

The procedure is shown on figure 6.1. With only one DSP, the advantage of a RTOS is only that it organises the calling structure. If there were no prospects of an expanded system version, then this RTOS would probably be too much work to use.

6.3 Purchasing

The purchasing of the hardware and software has been a part of the work carried out. Special conditions with purchasing expensive material has affected the project in such a way, that it deserves a description.

Purchase Procedure

At CERN all buys goes through the CERN purchasing service. When a request for a buy is done, there is a number of persons that has to approve of this, before it reaches this purchasing service.

An internal purchase request is done by a program called DAI[4]. This is a form, on the local network, which is filled out with information of the buy, such as price, budget source, distributor, delivery details, tax information[5], obtained discount etc. If it is an expensive buy(more than $\sim$ 1000 \$), then some additional documents should follow the request. It has to be verified that the best distributor and product is found and preferable bought from a member state country. Offers from up to 5 distributors has to be obtained for comparison reasons, if possible. In any case expensive or non expensive, an official offer must follow the internal purchase request.

Then it has to be approved by around 6-8 persons, before it is sent to the purchasing office. Here the request stays for a while before it is sent to the distributor. Only when the merchandise is received, the money is sent to the company.

This series of approvals takes a while, so offers may from time to time expire. This is what happened to us when we bought the hardware. The

[4]Demande d'Achat Interne

[5]CERN are not obliged to pay VAT

Product	Company	price in dollars
Pentek 6441 ADC	VSYSTEMS	8.077 $
Pentek 6510 DRX	VSYSTEMS	14.700 $
HW Emulator	Sonepar Electronique	1.485 $
Assembler,C-compiler,Linker	Sonepar Electronique	557 $
TMD320C4x Simulator	Sonepar Electronique	278 $
GO-DSP Code Composer	Sonepar Electronique	1.783 $
Diamond RTOS for C4x	3L Ltd.	1.113 $

Table 6.2: Purchased goods

software buy alone took 1 month from the day the request was made till the software was shipped. The hardware was delayed even more and moreover there was a bug on the board that had to be fixed by Pentek before shipment. All in all this resulted in a delay of close to 4 month. The hardware was received 1.10.1998. the request for buying it was made 25.5.1998.

Online Information

When the request is done it is possible to follow how long the request is on its way. Through the local network all approvals and information of what is done, with the request, is available. At each moment, you can see where the request awaits actions. All information is stored, so the history of the request can be inspected as well.

6.3.1 Purchased material

In direct relation to this project the following material is purchased. All prices are without VAT[6], because CERN is not obliged to pay this, as it is an international organisation. The currency conversion is done by the CERN currency converter on the WEB[7] the 15th of September 1998.

The purchased goods in table 6.3.1 are summed up to about 28.000 $, but this is only the first buy. When the equipment has proven its worths, the second 6441 ADC is bought and spares of the the whole system is ordered. At the end this sums up to around 67.000 $. Additionally specific equipment is made at the CERN location, to cope with the special demands this system requires. This material is f.ex. pick-ups, head amplifiers, attenuators, direct digital synthesisers etc. This is all equipment that is not commercially available, in the required shape.

[6]Value Added Tax

[7]http://cadd.cern.ch/eucwww/draft/exchange.html

6.3.2 Search of distributors

The distributor of embedded VME crate systems was present at the Syscomm98 conference, held at CERN. Personal contact was established here to M. Emmanuel THIBAUT from VSYSTEMS which became our provider of embedded modules for this system.

When it came to development software, we have searched the WEB for possible solutions and fixed our minds to the products we wanted. I received offers from 5 distributors[8] of these products. I was able to get university discount on two of those offers, which was a difference of ∼25%. Some of the TI products even had 40% discount for universities, so the remarkable discount was worth searching for. The university discount was obtained because of the fact that CERN is a non-commercial organisation. The only product leaving CERN is research results.

The RTOS we bought directly from the english company, 3L Ltd, who developed it. There were some problems buying this software as they didn't reply our emails or faxes. After a while we enabled contact and received the required official offer to make an internal request.

[8]EBV, VSYSTEMS, DENIMEX, AVNET EMG. S.A., Sonepar Electronique

Chapter 7

System software

The system software is the outcome of this project. The result of all the analysis done in the preceding chapters. The software is not yet fully developed, as this section will explain, but a restricted model is present and explained.

This chapter starts out with a light introduction to the current software version. Following is a section explaining each software block. The full detailed versions are referred to, in the appendices, of each description. At the end of the chapter, future software changes are mentioned. They are split in a detailed immediate task list, which is going to be carried out in the near future. Another less detailed future task list, is mentioned, but on a larger time scale.

7.1 Software structure

A sketch of the software structure is shown on figure 7.1. This is a state-machine-like drawing, of the current software version. This version program works as drawn, but has some restrictions which will be covered in this and later sections.

7.1.1 Overview of structure

On the sketch at figure 7.1 some rings are gray shaded, these are ISRs[1]. From almost anywhere in the program, it is possible to jump to these states. The only exception is during setup of hardware, where the interrupt environment is not yet enabled.

The software is fully interrupt controlled. After setup of hardware the program is put into an infinite loop, in which the program will stay if no interrupts occur. In this constant loop two conditions are checked. These conditions can only be changed due to interrupts. The two conditions

[1] Interrupt Service Routines

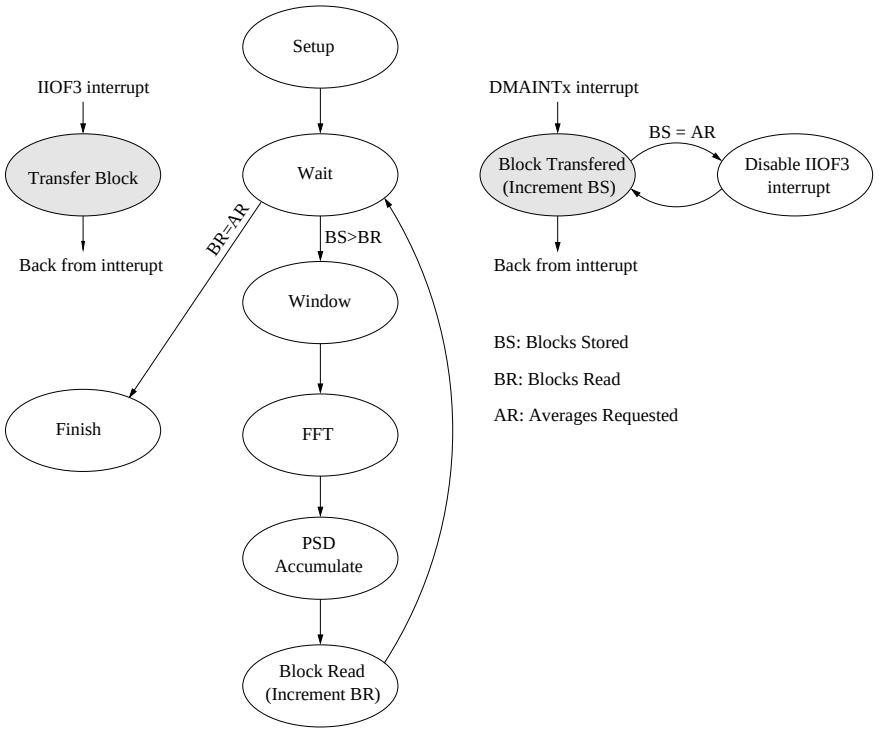


Figure 7.1: Overall software structure

checked can, if they become true, lead to processing of a data block or termination of the program. The processing condition requires a new data block, of unprocessed data. When such a new data block is received by the FIFO buffers, an ISR orders a transfer of data. When these data are transfered another ISR is activated. The second ISR changes the conditions (increment of BS), such that the loop is broken and data gets processed. This changes again the conditions (increment of BR).

The current version handles only one single block of 256 complex data. In principle this structure could handle several blocks without problems. The program already contains increment functions of, for example, the next block address. But the placement of Fourier transformed data requires bit reversal. The bitreversal with TMS320C40 instructions, as the current algorithm uses, is only possible with a zero offset base address. This is the main reason why the system is not yet expanded, to contain more than one block. In later sections an approach to this problem will be taken.

Enabling the software to process more than 512 complex samples requires two half-full FIFOs, thus two block transfers for one processing. This requires just simple logic, but due to unsolved problems with subsequent FIFO interrupts, this is not yet done.

7.2 Module descriptions

All of the modules uses global variables. As the software execution is based on interrupts, it is not possible to pass parameters by calls, in a normal fashion. In stead, some addresses are globally defined and the contents of these addresses, are modified by the software modules. These addresses are not situated in areas where any software is declared. Data is loaded into these addresses from a separate file, simulating that it is the DSC, which stores data.

The subsequent text is the declarations of addresses taken directly from the main function, of the system software. All the current variables are listed with their address location.

```
int     NextStore =     0x40000000;   /* address to store next dma*/
int     BlocksStored =  0x40000001;   /* the number of data blocks stored */
int     NextRead =      0x40000002;   /* address to read next data block */
int     BlocksRead =    0x40000003;   /* the number of data blocks read */
int     AccuStore =     0x40000004;   /* the address where the PSD are accumulated */
int     Averages =      0x40000005;   /* the number of averages requested */
int     FftSize =       0x40000006;   /* the fft size wanted */
int     LogFft =        0x40000007;   /* the logorithm(log2) of fft_size */
int     FifoAddress =   0x40000008;   /* the FIFO address */
int     DmaAddress =    0x40000009;   /* the address to put dma request */
int     DmaCtrl =       0x4000000A;   /* the DMA control word */
int     EIIOFxOn =      0x4000000B;   /* enable word for DMA channel */
int     EIIOFxOff =     0x4000000C;   /* disable word for DMA channel */
int     EDMAINTx =      0x4000000D;   /* word to set DMA interrupt enable bit */
int     DmaChannel =    0x4000000E;   /* the DMA channel used */
```

In assembly code, such variables has to be copied to the data segment of the code, where it is used. So a function written in the *.text* segment has its *.data* segment close to it. This is because the instruction words, are only 32-bit long and a displacement of a variable can not exceed the representation of 16 bits ($\pm$32.768 memory cells). So when the content of a global variable is used, it has to be done in three steps, example following.

- `LDI  @NextStore,AR5  ; AR5 loaded with address in local data segment`
- `LDI  *AR5,AR1        ; AR1 loaded with global NextStore address`
- `LDI  *AR1,R0         ; R0 with the NextStore parameter`

So this rather bizarre pointer-to-pointer-to-parameter, is needed in assembly order to load parameters far away from the *.text* segment. This instruction pattern, is seen in almost every assembly module.

The main function of the program contains a lot of settings of the DRX. These functions are not covered here and the main function source code is not enclosed to this report either. The only thing important for this program to execute right, is declaration of the above mentioned variables, execution *set_ivtp* and then afterwards *wait*. The rest will be done by interrupts or calls from *wait* due to condition flags. The structure of main is thus simply.

```
extern    void    set_itvp ();
extern    void    wait ();

Global variable declarations

main
{
          Setup DRX
          set_itvp ();
          wait ();
}
```

Common for all modules, is that they preserve the registers used. The extended register used for floating point representation, does not fit in a single 32-bit memory cell. They have to be PUSHed in two stages. First the mantissa, by PUSH, and then the exponent, by PUSHF. Only registers which is used in modules are preserved by them. To avoid significant overlap, reuse of registers is advised. The preservation of registers is very important for interrupt processing, as the interrupt can occur in the middle of an algorithm and has to restart it afterwards, without any loss of information.

Setup of interrupt environment

There are three things that has to be done, to setup an interrupt environment.

- First, the interrupt type has to be enabled. This is done by modifying either the IIF[2] or the IIE[3] register. For external interrupts on IIOF0-IIOF3 the lower 16 bits of IIF register has to be modified. We set the IIOF3 interrupt pin to external interrupts by putting a 1001_b on the 4 bits reserved for the IIOF3 interrupt. For internal interrupts, from the DMA, we need to set the corresponding bit in the IIE register. We also has to include an interrupt request in the DMA control word set by the Transfer Block ISR.
- Second, the IVT[4] has to contain addresses where the program is suppose to jump to, when an enabled interrupt occurs. This table is put

[2]Interrupt flag
[3]Internal Interrupt Enable
[4]Interrupt Vector Table

at a user defined place and the IVTP[5] holds the base address of the table. Only two memory cells are set, as we have only enabled two interrupts. The structure of the interrupt environment setup, is flexible regarding the DMA channel. Any of the six available channels, can be chosen from global variables.

- Third, interrupts has to globally enabled by setting the GIE bit in the ST[6] register to 1. Whenever an interrupt occurs, the GIE is set to 0 and PGIE to 1. This has the effect, that interrupts are disabled during execution of an ISR. As the FIFO half full interrupt is important, we set the GIE back to one in the *DMAINTx* ISR, which makes this ISR interruptible. This is fully legal. If such two interrupts has to be serviced at the same time, then the IIOF3 has higher priority.

The entire source code, for this assembly module, is written in appendix F

Wait

The *Wait* module is a simple module, it check continuously two conditions.

- If the number of processed data, at address *BlocksRead*, is equal to the requested found at address Averages. If it is, then it jumps calls the *finish* module.
- If the number of data blocks stored at address *BlocksStored* is bigger than what is processed, then a new block is ready to be processed. The current version only works for FFT lengths of 256.

The entire source code, for this assembly module, is written in appendix J

7.2.1 IIOF3 interrupt

The IIOF3 flag goes high, when a FIFO is half full. This was ordered in the DRX setup and set in the setup environment by the *set_itvp* function. When this flag goes high, the FIFO buffers need to be emptied. This is done by a DMA transfer request. This module passes the global parameters, for such a transfer by the stack and calls the *dma* transfer function. When the program execution returns from the *dma* function call, it puts back registers and returns from the interrupt.

The entire source code, for this assembly module, is written in appendix G

[5]Interrupt Vector Table Pointer
[6]Status

DMA transfer request

The *DMA* takes care of the data transfer, from FIFO buffers to 'C40 memory. This is done without disturbing the 'C40 code execution. Only for a short period, the DMA function posses the processing power. Only the time to transfer 7 values and then the DMA co-processor will process, on its own. For more information about the DMA, see page 77.

There are 6 DMA channels to chose from, their base addresses are:

dma-ch0 = 0x001000A0
dma-ch1 = 0x001000B0
dma-ch2 = 0x001000C0
dma-ch3 = 0x001000D0
dma-ch4 = 0x001000E0
dma-ch5 = 0x001000F0

The choice of channel is decided by the externally given word at *DmaAddress*.
The entire source code, for this assembly module, is written in appendix I

Disable IIOF3 interrupts

Disabling the IIOF3 interrupts is done by setting the 16th-19th bits in the IIF register to zero. This makes the CPU ignore the IIOF3 internal and external interrupts. These four bits has the following meaning.

- bit 19 (EIIOFx): Enable external interrupt (1)
- bit 18 (FLAGx): Interrupt asserted (1)
- bit 17 (TYPEx): Edge-triggered(0)/level-triggered(1)
- bit 16 (FUNCx): General purpose I/O pin(0)/Interrupt pin(1)

These are all R/W[7] bits, so an artificial interrupt can be created by asserting the FLAGx pin from debugger or software. All other interrupts are not affected by this modification. These four bits are masked and set to zero all other bit stays the same.

The entire source code, for this assembly module, is written in appendix G

[7] Read/Write

Hanning window

The windowing is done by a Hanning window. The coefficients are compiled as source code. The format of the file is:

```
      .globl   _taps256
      .sect    ".text
_taps256
      .float   1.49421078e-004
      .float   5.97595007e-004
      ....
      ....
      .float   5.97595007e-004
      .float   1.49421078e-004
      .end
```

There are two such files one for each FFT length, their global variables must be respectively *_taps256* and *_taps512*.

The data stored from FIFO buffers are represented in 16-bit 2's complement numbers. They are situated in the upper 16 MSBs of the 32-bit memory cells. To transform this 16-bit number to a 32-bit representation, the 16 LSBs are simply given the same value as the sign bit of each number. This happens when the values are read into registers. There are two values read in, at the same time. These are the real and imaginary part of a number. They both need to be multiplied with the same tap coefficient. The taps are, however, represented in floating point numbers, so we need to convert the 2's complement numbers into their floating point representation, before multiplying.

The entire source code, for this assembly module, is written in appendix L

FFT algorithm

The FFT algorithm was originally written by Texas Instruments. It was taken from their homepage at [21]. There were two errors, which were corrected by the author of this report. They were due to misuse of registers in parallel instructions. All code added to the original version has a !! symbol at the beginning of the comment line.

A lot more lines had to be changed, to adjust the algorithm to the specific use. The changes are due to:

- Different calling conventions than original version
  The calling conventions are described in the beginning of this section and they are simple to implement.

- Different FFT lengths(256/512) with use of single COSINE tap table (640 taps)
  The reuse of the same cosine tap table, requires some extra logic

deciding the step index and the displaced pointer to the cos base address.

The entire source code, for this assembly module, is written in appendix M

PSD and accumulation of results

The PSD density function is proportional to the squared absolute Fourier components. The complex values of the Fourier transformed signal, is thus squared and added to cells pointed to, by the destination pointer. Parameters are, as always, taken from the stored global variables.

The destination address space, has to be initialised by load in of a zero data table, no initialisation is done by the program itself.
The entire source code, for this assembly module, is written in appendix N

Increment of BR

The cell for blocks read is a simple straight forward module. Its description do not need further explication
The entire source code, for this assembly module, is written in appendix O.

7.3 Software changes

This model of the software is a small version, of what is foreseen. Due to late arrival[8] of the hardware, the writing software only lead to this restricted version. A lot of changes are, however, foreseen and the most clear of them is described in this section.

7.3.1 Immediate software changes

The immediate expansion, of the software, is aimed to process a request on a single data channel. This requires the following changes.

Bitreversed addressing

The bit reversed addressing is not possible to do afterwards, in subsequent blocks of memory. This is due to the TMS320C40 instruction set, that requires a zero offset base address, to bitreverse. Another approach must be taken. There are several options, either the bit-reversing is done before the processing and the data is stored in subsequent blocks. Or the bit-reversing is still done after processing, but an arrangement of zero offset base address blocks, are reserved. Then afterwards data is moved around. No approach is chosen, yet.

[8] Approximately six weeks before delivery of this report

FIFO buffers

The FIFO buffer interrupts are not fully controlled. Multiple interrupts from FIFOs fails to appear. Only after FIFO reset, it is possible to get an interrupt. The FIFO buffer inspection, is impeded by the fact, that it is not possible to freeze the FIFO memory. It is either all zeros or data is shifted in at minimum 300 Hz.

Expansion of global variables

The global variables are going to be expanded in number. They should contain every possible setting of the software, including the DRX setup. This is not entirely done by only the author of this report, but involves the composer of the DRX software setting, as well. Around the end of November, a new data system draft should be ready for modification and approval.

Profiling and optimising

The profiler is not set up right or is not working, as it is explained in the Code Composer manual. The current version calculates processing loads, about 10 times those, calculated by the simulator. In one way or another, the processing load has to be analysed. If the processing does not meet the real-time constraints, then further code optimising, has to take place.

7.3.2 Future software changes

The future software changes requires a lot more programming than just assumed at first sight. The tasks written beneath, will probably involve at least two-three man-month of programming, just to be positive.

Multiple data channels

When one single channel is working, the system is expanded to, the current maximum of two input channels. The system can still, activate 4 for digital receiver chips. Thus, maximum system data rate simulation is possible.

RTOS

When an expanded version done and working satisfactory, this version is saved. In case further development is delayed, this version fulfills simple urgent needs. A final version with the RTOS, can begin. This involves slight changes of software modules. The change should be limited, as the former structure had this RTOS in mind. It is, however, a modification of all of the software written, so some time for this transition is reserved.

Final crate implementation

When program package is fully operational and debugged, it is ready for implementation in the real VME crate in the control room. No complications or significant work is foreseen for this task.

DSC software

The DSC software is going to be composed by the aid of the system definition paper. In case of complications or unclear points, this will need my presence as system acquainted. As mentioned earlier, this will take place in January and February 1999 by Jean-Henry Hemelsoet.

Workstation software

The workstation software is written at a later stage of the project. When the lower system programs are working properly then software development, at this level, can begin.

Chapter 8

Performance

8.1 Current system state

The current system software, is covered in chapter 7. Figure 8.1 is a printout of the files window in Code Composer. This is a good image, of the current state of the system[1]. All of the modules, used for the current version, are shown. The main.c and init6510.c are used to setup the digital receiver. They use the include files listed in the include directory. All of these aren't covered in details here and they are too long to enclose as appendix.

This section covers the performance of the equipment, with these files compiled and downloaded into DSP memory.

The task being able to compile, download and run these few software modules, is however more comprehensive than writing them. During this project, a complete development environment has been setup from scratch. This environment comprises:

- 1 Pentek 6441 ADC card
- 1 Pentek 6510 DRX card
- 1 STR 700 VME crate with power supply and ventilator
- 1 3M 80-pin high density flat ribbon cable (ADC-DRX connection)
- 1 Texas Instruments J-Tag cable
- 1 J-Tag emulator card, for PC
- 1 486 Intel PC running Code Composer for Windows and has the TI compiler package installed
- 3L Diamond RTOS for TMS320C40

[1] November 1998

Figure 8.1: The software files currently written

- (Stanford Research Systems DS345 Signal generator)
- (Tektronix TDS 320 Oscilloscope)

The code composition, debugging and testing is done directly in this environment, in the laboratory.

8.2 Current system performance

The system is only just setup approximately six weeks before this report was handed in. In this period, a lot of tasks needed to be carried out. Before the real coding could begin, quite some time was wasted before we got a "hole through", to the DSP. This has had an effect on the verification phase. Due to time constraints, it has not been possible, to carry out as profound as one could wish.

Some material is however present and presented in this section. It should be noted, that this section is not in any way adequate, to fully describe equipment and software performances.

8.2.1 Envelope function test

The Harris chip uses a composite filter, consisting of two serial filters. The characteristics, taken from the manual, is shown at the left hand side of

figure 8.2. At the right hand side, the measurement of the real performance is shown. Wide band noise is generated by the signal generator and connected to the input of the ADC. The downconverted signal is Fourier transformed and the result shown at the right hand side. These two graphs should correspond. From just graphical inspection of the figures, we can verify this.

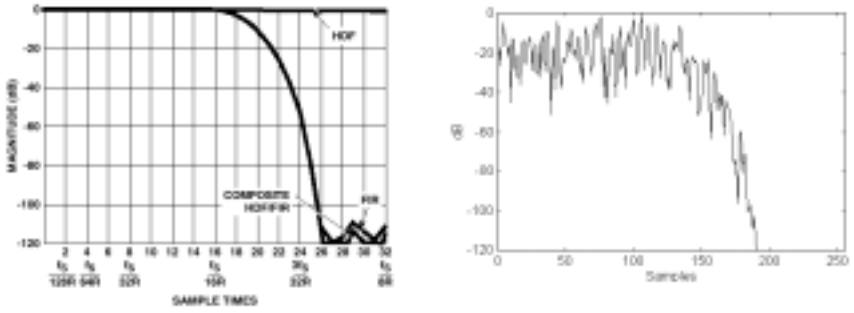


Figure 8.2: Left: Data sheet envelope Right: Measured wide band noise

Another test, with generated sine waves at different frequencies, has been performed. The LO frequency was set to 6 MHz and the decimation rate to 223. From the Pentek manual, they promise that we should be able to see a frequency range of:

$$\Delta f = \pm 0.66 \frac{40MHz}{2 * 223} = \pm 59kHz$$

- with less than 3 dB of damping. The results of varying the input frequency in different tests should reveal the envelope function. The results was as follows.

frequency	peak at bin number	amplitude
6 MHz	0	∼9.8e21
5.99 MHz	14	∼9.9e21
5.97 MHz	42	∼9.8e21
5.95 MHz	70	∼5.5e21
5.94 MHz	84	∼0.14e21
5.93 MHz	no peak	max. 2.5e15

From this it can be concluded that the Pentek manual number of 59 kHz frequency range, is a little bit too much. This number couldn't be found in the Harris chip manual, neither. The results shows, however, good accordance with envelope of the filters mentioned in the Harris chip manual. For future reference, the Harris chip manual is used.

8.2.2 Downconverted data

After complex downconversion of input data, a filtering is done by a Hanning window. The signal is, after such treatment, shown on figure 8.3. This is a 5.99 MHz sine wave, which is downconverted with an LO frequency of 6 MHz. The decimation rate was set to 223.

It is seen, that the real versus imaginary data are exactly 90 degrees out of phase, as they should. The Hanning window is easily recognised, from the envelope of the oscillations.

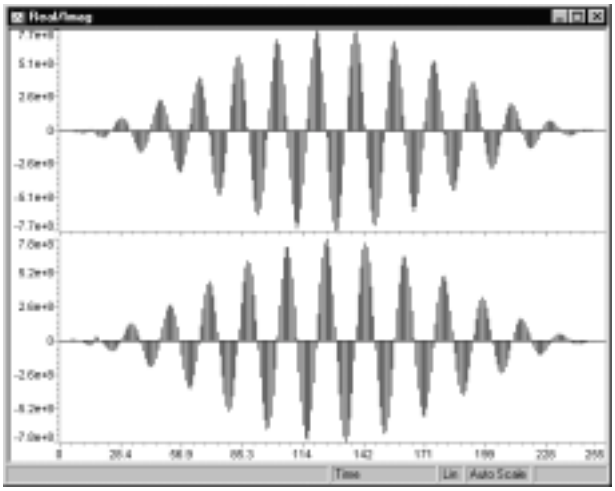


Figure 8.3: The downconversion of 5.99 MHz with an LO frequency of 6.00 MHz

This shows that both Hanning window filtering and the real and complex part of the downconverted signal are downconverted right and put correctly into 'C40 32-bit memory. Just by looking at this nice sine wave, we know that:

- The setup of the DRX seems to be done properly. High frequency downconverted to a low and no immediate malfunctioning.
- The interrupt setup and interrupt routines are well coordinated. If not none of the function calls would have been made.
- The FIFOs are emptied properly by the DMA.
- The Hanning window filtering is working and the 16 bit output data are properly converted to 32 bit representation.

8.2.3 Fast Fourier Transformation

The FFT algorithm was run on a simulator, before the hardware arrived. In this environment, the functionality of the algorithm was verified. To

test that the version downloaded into DSP memory worked the same way, a printout of data from the simulator was used. As mentioned earlier, it is not easy to export data from the simulator, it has to be done by manually by a cut-and-paste fashion. Input data was produced in Matlab and enclosed as program code, in both environments.

The two implementations corresponded.

8.2.4 Processing of a signal

We should be able to calculate the intensity from the Fourier transformation. The following tests, analyses how exact the processing software in the 'C40 does this task.

A ± 1 Volt oscillating signal was connected to the ADC input and this resulted in a downconverted cosinewave of amplitude:

$$A_{cosine} = 1.569947e9\,V$$

This oscillation can theoretically be calculated to posses an energy of:

$$P = A_{cosine}^2 = 24.64734e17\,W$$

But the input is filtered by a Hanning window, so the power decreases. According to the calculations made in section 3.5, we can calculate the new signal power as:

$$P = 24.64734e17 * 0.3757324 = 9.260804e17\,W$$

The signal is then Fourier transformed and squared by the 'C40 software. A file of data is stored and summed up in Matlab. This summed up value of the processed data is:

$$\hat{P} = 9.252553e17\,W$$

Thus a normalised error of $0.89e-3 \approx 1\%$ compared to the theoretical value. Note that $\hat{P}$ is less than P because of the bias calculated in section 5.3.3 on page 93.

8.2.5 Processing load

As mentioned in former sections, the profiling feature of Code Composer has not been made to work. From simulations of the most time consuming module, the FFT, the processing loads in table 8.1 are obtained.

Table 8.1: FFT processing load

FFT length	clock cycles	time consumption
256	26.697	0.533 msec
512	59.195	1.184 msec

Chapter 9

Conclusion

This project has meet its overall project scope. A system has been designed and software, for analysis of beam parameters, has been developed. All of the required phases, for high performance data acquisition and processing system development, has been gone through. The origin of problem, has been well defined and the project covers, how this problem can be measured. The development environment, for an embedded VME crate based system, is put together. Software for a laboratory model has been written and works as a minor version, of the final data acquisition and processing system. Sparse testing of equipment and composed software is carried out.

The different phases of the development are however influenced by the late arrival of equipment. This has affected the last development phases, significantly. The consequence of this, is a small model of the system software and lack of testing. The different software modules are only just put together, in a way that it resembles the final version. The final version, will undoubtedly take another different shape.

Another regrettable consequence, is that calculations being build up during the project, has not yet been applied. They should have been verified in a missing extended testing phase. The preliminary work to further development, is however well founded. Even though the final version of the system is not obtained, most of the work done in this project, will be recognised when such a system is ready.

Kristian Philip JØRGENSEN

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Appendix A

Explaination of EXCEL timing sheet

`p from:` The momentum at the beginning of the time slice. Values taken from AD cycle specifications

`p to:` The momentum at the end of the time slice. Values taken from AD cycle specifications

`f_rev from:` The revolution frequency at the beginning of the time slice. The frequency is calculated from

$$f_r = p\frac{1}{L}\sqrt{\frac{c^2p^2}{c^2m_o^2 + p^2}}$$

This formula was introduced in subsection 2.4.4.

`f_rev to:` The revolution frequency at the end of the time slice. Calculated as written above.

`K_i to:` The ratio between revolution frequency and sampling frequency.

`f_s from:` The sampling frequency at the beginning of the time-slice.

$$f_s = k + i * f_{rev}$$

`f_s to:` The sampling frequency at the end of the time-slice.

`R:` The decimation ratio. The ratio of which the incoming samples are reduced

`BW/f_rev:` The bandwidth-revolution-frequency ratio.

$$\frac{BW}{f_{rev}} = 0.6\frac{f_s}{4Rf_{rev}}$$

`Harmonic + sign of q:` The harmonic situated in the band between 5.5 and 6.5 to be zoomed upon.

`LO from:` The local oscillator frequency at the beginning of the time-slice. This should be less than 6.5 MHz.

$$f_{LO} = f_{rev}(m \pm q)$$

Formula introduced in this very section.

`LO to:` The local oscillator frequency at the end of the time-slice.

`F/2`hat`32:` The ratio between the local oscillator frequency and the sampling frequency.

`N_av:` The number of averages done of spectra.

`N_fft:` The size of the fft.

`meas. time:` The time for carrying out the averaging of the elegit number of fft's. This should be just a "flash" of low duration.

$$T_{meas} = N_{fft}\frac{N_{av}+1}{2}f_{sc}$$

`# of meas.:` The number of measurements possible in the time-slice.

$$N = \frac{T_{period}}{T_{meas}}$$

`delta_q/bin:` The resolution of which the q value can be distinguished.

$$\Delta q = \frac{BW}{0.6N_{fft}f_{rev}}$$

`total period:`	The total period of the time-slice.
`complex sample rate (highest):`	The rate that the samples leave the DRX to enter the DSP at the beginning of the time-slice.
`complex sample rate (lowest):`	The rate that the samples leave the DRX to enter the DSP at the end of the time-slice.
`angle of envelope(START):`	The angle of the envelope function at the beginning of the time slice. $$angle = 360\frac{5.5}{182 f_{rev}}$$
`angle of envelope(END):`	The angle of the envelope function at the end of the time slice. $$angle = 360\frac{6.5}{182 f_{rev}}$$

Appendix B

Specifications for Pentek 6441 ADC

Input Single ended	$\pm 1[V]$ full scale $50\,[\Omega]$ input impedance
Anti-aliasing filter	DC to $16\,[MHz]$ $\pm 1\,dB$ passband flatness $24\,[MHz]$ stopband with $> 50\,dB$ attenuation (bypassed by jumper)
A/D converter	$12\,bits, 41\,[MHz]$ max. sampling rate Analog Devices chip (AD 9042) SNR$> 65\,dB$, SINAD$>$ 60 dB spurious components $< -80\,dB$
Sampling clock	internal $40\,[MHz]$ crystal oscillator, or user installable DIP TTL oscillator external TTL clock through front panel SMA connector
Power	$1.0\,[A]$ at $+5\,[V]$ $0.75\,[A]$ at $+12\,[V]$ $1.0\,[A]$ at $-12\,[V]$
Size	6U board 6.3 [in] *times* 9.2 [in], panel 0,8 [in] wide

Appendix C

Specifications for Pentek 6510 DRX

Receiver type	Harris HSP50016
Digital input format	four independent inputs each input with 16-bit words 2's complement one sample clock line
Input level	TTL single ended
Sampling rate	DC to 50 [MHz] max.
Data input connector	80-pin flat ribbon cable 0.025" pitch (3M)
Input multiplexers	two groups of four receiver channels each each group can independently select one of two front panel inputs under program control
Local oscillator	direct digital synthesizer single frequency CW and sweep (chirp) up/down modes CW frequency is equal to $F * f_s/2^{32}$, where F is 32 bit binary integer and f_s is the input sample rate

Tuning range	DC to $f_s/2$ (25[MHz] for $f_s = 70[MHz]$)
Tuning resolution	$fs/2^{32}$ ($\sim 0.008[Hz]$ for $f_s = 50[MHz]$)
Low pass filter	decimating 121-tap FIR programmed by 15-bit integer R, from 16 to 32768, nominal output Nyquist bandwidth $f_N = f_s/4R$, output sampling rate is $f_s/4R$ for complex outputs and $f_s/2R$ for real.
Filter response	$\pm 0.04dB$ ripple bandwidth $= 0.6f_N$ $-3dB$ bandwidth $= 0.66f_N$ $-105dB$ stop bandwidth $= f_N$
Real mode	16-bit real output samples at sampling rate $f_s/2R$
Complex mode	16-bit complex (Interleaved I and Q) output samples at sampling rate $f_s/4R$ per complex pair

Appendix D

TMS320C40 block diagram

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Figure D.1: 'C40 Block diagram (1 of 2)

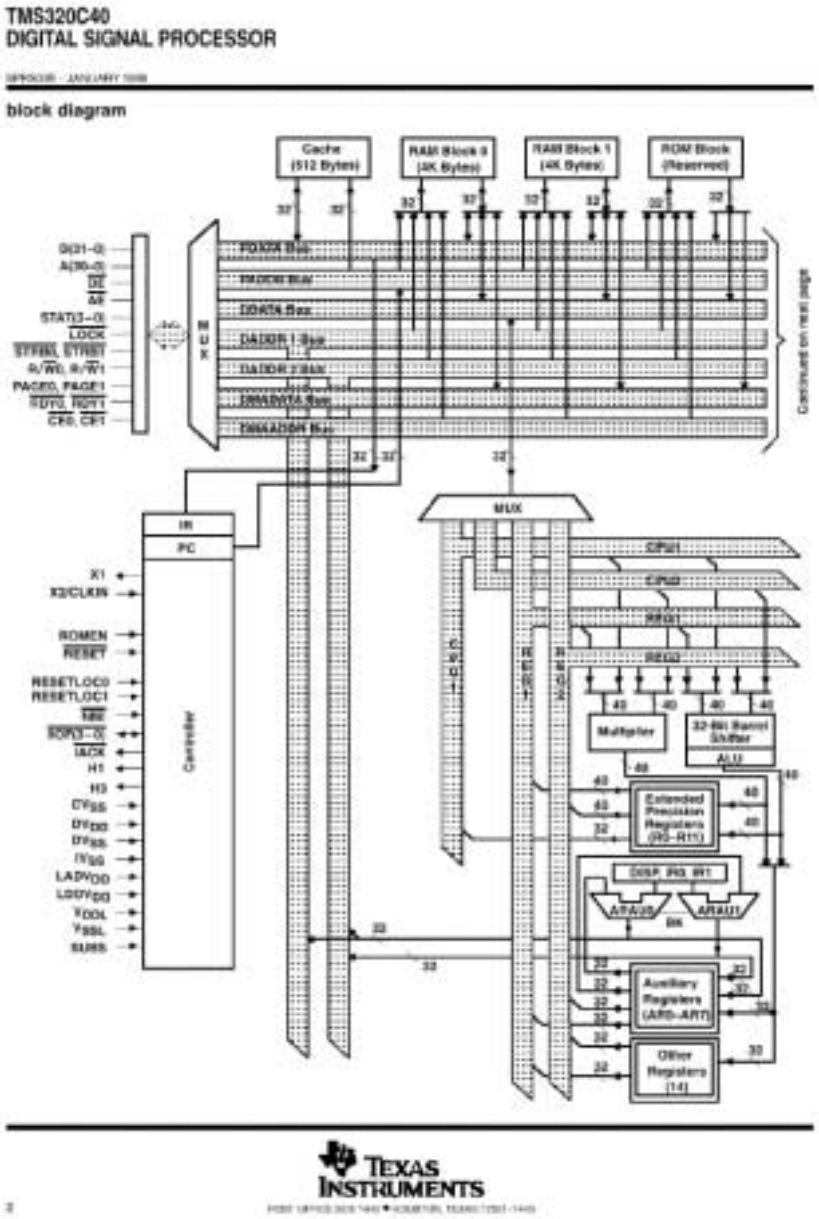
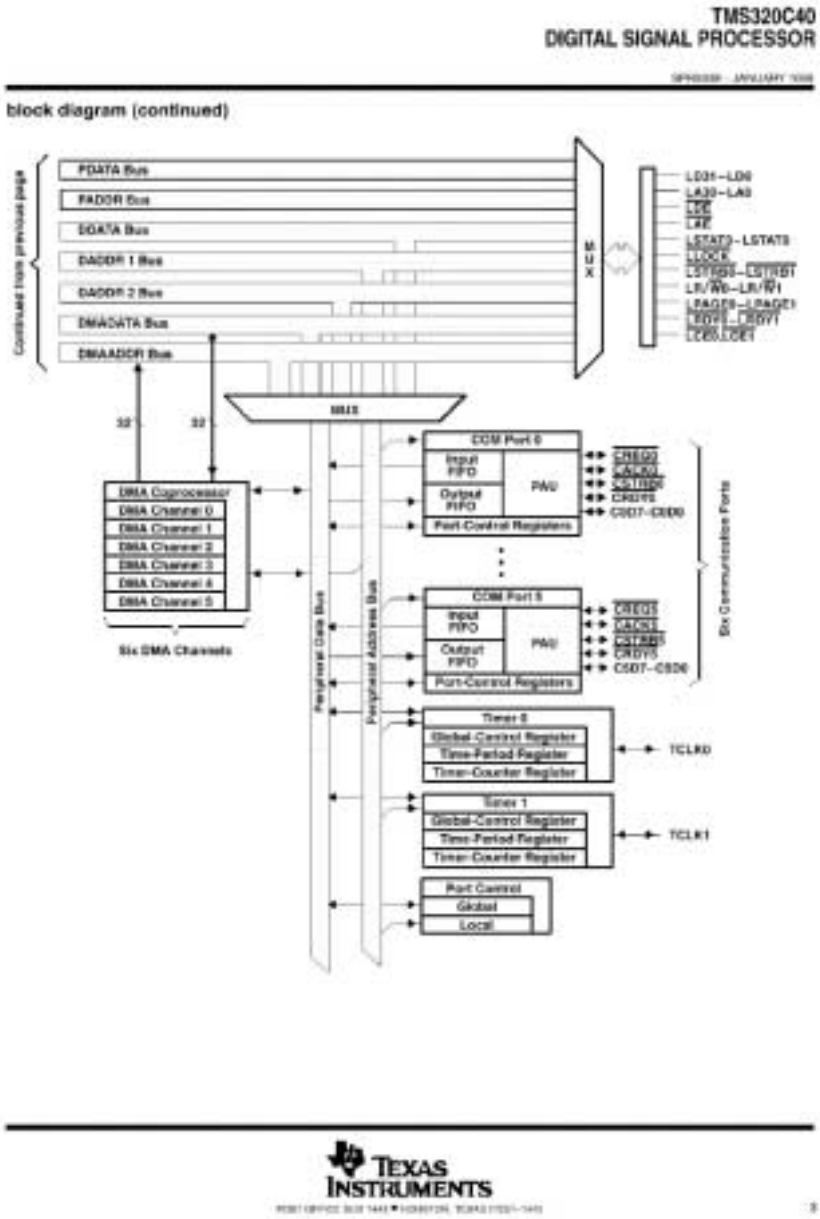


Figure D.2: 'C40 Block diagram (2 of 2)



Appendix E

HP48G/GX program for TI floating point conversion

```
≪
DUP
IF # 80000000h ==
THEN DROP 0
ELSE 32 STWS DUP
HEX # FF000000h AND
SRB SRB SRB 8 STWS
DUP B→R
IF 2 7 ^ ≥
THEN NEG B→R NEG
ELSE B→R
END 25 STWS -23 + 2 SWAP ^ SWAP
# FFFFFFh AND DUP
B→R
IF 2 23 ^ ≥ THEN # 1800000h XOR NEG B→R NEG
ELSE # 800000h OR B→R
END ∗ END 32 STWS
≫
```

With this program you type for example $\#32F67CA9h$ and push the program name on the VAR menu and the program will replace the number on the HP48 stack with $-1.20957755E15$ which is the corresponding value of this TI floating point number.

Appendix F

Source code: set_ivtp.asm

```
*************************************************************************
*                        SETUP INTERRUPT ENVIRONMENT
*
*         Function written by Kristian Jorgensen PS/RF CERN  November 1998
*
*************************************************************************
*
*      This function is setting up the interrupt environment
*      meaning:
*
*      #1:            The IIF (interrupt flag) register
*
*      #2:            The IIE (internal interrupt enable) register
*
*      #3:            The ST (status) register GIE bit
*                      GIE = Global Interrupt Enable
*
*      #4:            The ITV (interrupt vector table)
*                      the table from where the CPU takes the next address
*                      when an interrupt occurs. The ITVP (interrupt vector
*                      table pointer) has to point at the base address.
*                      The ITV is shown and described at page 7-16 in the
*                      TMS320C40 User's manual
*
*      The table is placed at address table and the table+6h contains the
*      address of the ISR _data_transfer_order
*************************************************************************

          .globl          _EIIOFxOn
          .globl          _EIIOFxOff
          .globl          _EDMAINTx
          .globl          _set_ivtp
          .globl          _data_transfer_order
          .globl          _block_transfered
          .globl          _DmaChannel
          .sect           ".data"

table     .word            80001000h
F0value   .word            _EIIOFxOn
F0_mask   .word            _EIIOFxOff            ; IIOF3 set to external interrupt
INT3      .word            _EDMAINTx             ; word used set EDMAINT3 to one
STvalue   .word            00002000h             ; GIE position in ST word
ISR1      .word            _data_transfer_order
```

```
ISR2        .word           _block_transfered
DC          .word           _DmaChannel

            .sect     ".text"
_set_ivtp
            PUSH      AR0                       ; push used values
            PUSH      AR1
            PUSH      AR2
            PUSH      AR3
            PUSH      AR5
            PUSH      R1
            PUSH      R2
            PUSH      R3
            PUSH      R4
            PUSH      R5

            LDP       table                     ; load data pointer
* IIF register modification
            LDI       @table,AR0                ; load interrupt vector table base address
            LDI       @F0_mask,AR1
            LDI       *AR1,AR2
            LDI       *AR2,R1
            LDI       @F0value,AR1
            LDI       *AR1,AR2
            LDI       *AR2,R2
            LDI       IIF,R3
            AND       R1,R3                     ; Erase old F0 bits
            OR        R2,R3                     ; set new F0 bits
            LDI       R3,IIF                    ; set IIF register
* IIE register modification
            LDI       @INT3,AR1
            LDI       *AR1,AR2
            LDI       *AR2,R1
            LDI       IIE,R3
            OR        R1,R3                     ; set EDMAINT3 bit to 1, the rest unchanged
            LDI       R3,IIE

* Make interrupt vector table
            LDPE      AR0,IVTP                  ; set interrupt vector table base address
            LDI       @ISR1,AR1                 ; load interrupt vector for IIOF0
            LDI       @ISR2,AR2
            LDI       @DC,AR5
            LDI       *AR5,AR3
            LDI       *AR3,IR0
            ADDI      25h,IR0                   ; offset DMA interrupt = 25h
            STI       AR1,*+AR0(6h)             ; set IIOF0 interrupt vector
            STI       AR2,*+AR0(IR0)            ; set DMA transfer done interrupt vector

* ST register modification
            LDI       @STvalue,R4
            LDI       ST,R3
            OR        R4,R3                     ; set GIE bit to 1, the rest unchanged
            LDI       R3,ST


            POP       R5
            POP       R4
            POP       R3
            POP       R2
            POP       R1
            POP       AR5
            POP       AR3
```

```
POP       AR2                         ; pop used values
POP       AR1
POP       AR0
RETS
.end
```

Appendix G

Source code: iiof3.asm

```
*************************************************************************
*                      ISR ORDERING DMA BLOCK TRANSFER
*
*       Function written by Kristian Jorgensen PS/RF CERN  November 1998
*
*************************************************************************
*
* SYNOPSIS:     None (Not callable, must be implemented in interrupt vector)
*
*    No paramters needed, they are set globally
*
*    This ISR orders a DAM transfer from information in data header. Everything
*    is controlled by the data header: source, destination, DMA channel,
*    DMA control word, transfered values.
*
*************************************************************************

        .globl  _NextStore
        .globl  _FftSize
        .globl  _FifoAddress
        .globl  _DmaAddress
        .globl  _DmaCtrl
        .globl  _dma
        .globl  _disable_dma_int
        .globl  _data_transfer_order
        .sect   ".data"
NS      .word   _NextStore
FS      .word   _FftSize
FA      .word   _FifoAddress
DA      .word   _DmaAddress
DC      .word   _DmaCtrl
        .sect   ".text"
_data_transfer_order
        PUSH    DP
        PUSH    R4                  ; Save dedicated registers
        PUSHF   R4          PUSH    AR0
        PUSH    AR4
        PUSH    AR5
        PUSH    AR6
        PUSH    R8
        LDP     @NS

* Parameters put on the stack in reverse order
        LDI     SP,AR4              ; AR4 = virtuel stack pointer
```

```
ADDI    01,AR4
STI     1,*AR4              ; destination index increment
ADDI    01,AR4
LDI     @NS,AR5
LDI     *AR5,R4
STI     R4,*AR4             ; destination address
ADDI    01,AR4
LDI     @FS,AR5
LDI     *AR5,R4
STI     R4,*AR4             ; data transfer amount
ADDI    01,AR4
STI     0h,*AR4             ; source index increment
ADDI    01,AR4
LDI     @FA,AR5
LDI     *AR5,R4
STI     R4,*AR4             ; source address
ADDI    01,AR4
LDI     @DC,AR5
LDI     *AR5,R4
STI     R4,*AR4             ; control register
ADDI    01,AR4
LDI     @DA,AR5
LDI     *AR5,R4
STI     R4,*AR4             ; DMA address


ADDI    7h,SP
CALL    _dma
SUBI    7h,SP               ; compensate for stack incr. used for param. passing


POP     R8
POP     AR6                 ; Restore the register values and return
POP     AR5
POP     AR4
POP     AR0
POPF    R4
POP     R4
POP     DP
RETI
.end
```

Appendix H

Source code: dmaintx.asm

```
*************************************************************************
*                      ISR MODIFYING DATA HEADER
*
*       Function written by Kristian Jorgensen PS/RF CERN  November 1998
*
*************************************************************************
*
* SYNOPSIS:     None (Not callable, must be implemented in interrupt vector)
*
*    No paramters needed, they are set globally
*
*
*    This module is modyfying the data header cells:
*        BlocksStored and
*        NextStore
*    -if the    blocks stored is equal to the number of averages
*    requested, then a call to disable_dma_int in order to disable
*    the FIFO half full interrupt.
*
*************************************************************************


        .globl  _NextStore
        .globl  _BlocksStored
        .globl  _FftSize
        .globl  _block_transfered
        .globl  _Averages
        .globl  _disable_dma_int
        .sect   ".data"
NS      .word   _NextStore
BS      .word   _BlocksStored
FS      .word   _FftSize
A       .word   _Averages
STvalue .word   00002000h           ; GIE position in ST word
        .sect   ".text"
_block_transfered
        PUSH    DP
        PUSH    R0
        PUSHF   R0
        PUSH    R1
        PUSHF   R1
        PUSH    R3
        PUSHF   R3          PUSH    R4
        PUSHF   R4
```

```
        PUSH    AR1
        PUSH    AR4
        PUSH    AR5

        LDP     @NS

* set GIE = 1 which makes this ISR interruptable
        LDI     @STvalue,R4
        LDI     ST,R3
        OR      R4,R3               ; set GIE bit to 1, the rest unchanged
        LDI     R3,ST

* modify data header
        LDI     @BS,AR5             ; Blocks stored modified
        LDI     *AR5,AR1
        LDI     *AR1,R0
        ADDI    01h,R0              ; add 1 to Block Stored
        STI     R0,*AR1             ; R0 is also return value to c environment

        LDI     @FS,AR5
        LDI     *AR5,AR1
        LDI     *AR1,R1             ; load FFT size
        LSH     01h,R1              ; multiply FFT size by two
        LDI     @NS,AR5
        LDI     *AR5,AR1
        LDI     *AR1,R4             ; load Next Read
        ADDI    R1,R4               ; NextStore = NextStore + 2*FFT_size
        STI     R4,*AR1
* check if all blocks are transfered
        LDI     @A,AR5
        LDI     *AR5,AR1
        LDI     *AR1,R1

        CMPI    R0,R1
        CALLZ   _disable_dma_int



        POP     AR5
        POP     AR4
        POP     AR1
        POPF    R4
        POP     R4
        POPF    R3
        POP     R3
        POPF    R1
        POP     R1
        POPF    R0
        POP     R0
        POP     DP

        RETI
        .end
```

Appendix I

Source code: dma.asm

```
*************************************************************************
*                      DMA BLOCK TRANSFER FROM FIFO
*
*       Function written by Kristian Jorgensen PS/RF CERN  Oktober 1998
*
*************************************************************************
*
* SYNOPSIS:         dma( dmax , ctrlreg , source , s_index , transfer, dest , d_index )
* is loaded from stack to: AR2       R2
*
*        int        dmax:        dma channel base address
*        int        ctrlreg:    corntrol register
*        int     source:     source address
*        int        s_index:    increment counter of source addresses
*        int        transfer:    the number of memory cells to be transfered
*        int        dest:        destination address
*        int        d_index:    increment counter of destination addresses
*
*
* ASSEMBLY MODULE DESCRIPTION:
*    This module places a memory transfer request in dma memory. The base address
*    for the DMA channels can be one of the following:
*
*            dma-ch0 = 0x001000A0
*            dma-ch1 = 0x001000B0
*            dma-ch2 = 0x001000C0
*            dma-ch3 = 0x001000D0
*            dma-ch4 = 0x001000E0
*            dma-ch5 = 0x001000F0
*
*    The request for a transfer contains three 9 words, explained in Texas
*    Instruments User's guide chapter 11. The order of these words are the same as
*    in the dma function call. The control word are transfered as the last word.
*    When this word is placed, the transfer begins, if nobody else is using
*    the dma channel.
*    The control word is set from the function call but the advised word is:
*              0x00C40107
*    - the control word meaning can be found in the User's guide at page 11-8.
*
*    The address spaces of the different FIFOs are:
*
*        FIFO_1    0x80030000 - 0x80031FFF
*        FIFO_2    0x80032000 - 0x80033FFF
*        FIFO_3    0x80034000 - 0x80035FFF
```

```
*        FIFO_4    0x80036000 - 0x80037FFF
*        FIFO_5    0x80038000 - 0x80039FFF
*        FIFO_6    0x8003A000 - 0x8003BFFF
*        FIFO_7    0x8003C000 - 0x8003DFFF
*        FIFO_8    0x8003E000 - 0x8003FFFF
*
*    - the addresses from the VME bus can be found at section 3.5
*    in the 6510 manual.
*
*
*
*   Registers used (thus PUSHED/POPPED): R0,R1,R2,R3,R4,R5,AR0,AR1,AR4
*
*******************************************************************************

        .globl  _dma
        .sect   ".text"
_dma
        LDI     SP,AR0
        PUSH    DP
        PUSH    R0
        PUSHF   R0
        PUSH    R4                  ; Save dedicated registers
        PUSHF   R4
        PUSH    AR0
        PUSH    AR1
        PUSH    AR2
        PUSH    AR5



* read paramters from stack
        LDI     *-AR0(1),AR5        ; dma address
        LDI     *AR5,AR1


        LDI     *-AR0(3),AR5        ; source address
        LDI     *AR5,R4
        STI     R4,*+AR1(1)         ; set source address

        LDI     *-AR0(4),R4         ; source index increment
        STI     R4,*+AR1(2)         ; set source index increment

        LDI     *-AR0(5),AR5        ; transfer number
        LDI     *AR5,R4
        LSH     01h,R4              ; multiplied by two (because complex)
        STI     R4,*+AR1(3)         ; set transfer number

        LDI     *-AR0(6),AR5        ; destination address
        LDI     *AR5,R4
        STI     R4,*+AR1(4)         ; set destination address

        LDI     *-AR0(7),R4         ; destination index increment
        STI     R4,*+AR1(5)         ; set destination increment

* send DMA request
        LDI     *-AR0(2),AR2        ; control register
        LDI     *AR2,R0
        STI     R0,*AR1             ; set control word as last
```

```
END:    POP     AR5
        POP     AR2
        POP     AR1
        POP     AR0
        POPF    R4
        POP     R4
        POPF    R0
        POP     R0
        POP     DP
        RETS
        .end
```

Appendix J

Source code: wait.asm

```
*************************************************************************
*                        DUMMY WAIT MODULE
*
*         Function written by Kristian Jorgensen PS/RF CERN  November 1998
*
*************************************************************************
*
*      This module checks two conditions constantly:
*
*       #1: if BlocksRead are bigger than BlocksStored
*            then process data (process.c)
*
*      #2:      if BlocksRead are equal to Avereges requested
*            then finish process (finish.asm)
*
*      The module is sensitive to two interrupts:
*
*      #1:      IIOFx interrupt, an interrupt from the FIFO buffer x
*            when it is half-full and needs to be emptied
*
*      #2:      DMAINTx interrupt, an interrupt from the DMA when
*            a transfer is done.
*
*************************************************************************

            .globl      _Averages
            .globl      _BlocksRead
            .globl      _BlocksStored
            .globl      _FftSize
            .globl      _NextRead
            .globl      _LogFft
            .globl      _AccuStore
            .globl      _wait
            .globl      _finish
            .globl      _Process_data_block
            .sect      ".data"
A           .word      _Averages
BR          .word      _BlocksRead
BS          .word      _BlocksStored
FS          .word      _FftSize
NR          .word      _NextRead
AS          .word      _AccuStore
LF          .word      _LogFft
            .sect      ".text"
```

```
_wait
            PUSH      R1
            PUSH      R2
            PUSH      R3
            PUSH      R4
            PUSH      AR1
            PUSH      AR5

CHECK:      LDP       @A
            LDI       @A,AR5
            LDI       *AR5,AR1
            LDI       *AR1,R4
            LDI       @BR,AR5
            LDI       *AR5,AR1
            LDI       *AR1,R1
            LDI       @BS,AR5
            LDI       *AR5,AR1
            LDI       *AR1,R3
            LDI       @FS,AR5                        ; preparing of flexible fft length version,
            LDI       *AR5,AR1
            LDI       *AR1,R2

            CMPI      R4,R1                          ; checks if Block read are equal
            BZ        _finish


            CMPI      R1,R3

* pushing parameters on the stack
            LDI       @AS,AR5
            LDI       *AR5,AR1
            PUSH      AR1
            LDI       @LF,AR5
            LDI       *AR5,AR1
            PUSH      AR1
            LDI       @FS,AR5
            LDI       *AR5,AR1
            PUSH      AR1
            LDI       @NR,AR5
            LDI       *AR5,AR1
            PUSH      AR1

            CALLP     _Process_data_block            ; jump if BS>BR

* "virtual popping"

            SUBI      4h,SP

            BU        CHECK

            POP       AR5
            POP       AR1
            POP       R4
            POP       R3
            POP       R2
            POP       R1
```

Appendix K

Source code: process.c

```
/*
        PROCESSING OF DATA BLOCK (WINDOW/FFT/ACCUMULATE)

        Module wirtten by Kristian Jorgensen, November 1998

    This module processes a block of data found at the content of
    the NextRead. The data is Hanning windowed, Fourier transformed,
    squared and accumulated to former processed data. At last the content of
    BlocksRead is incremented.

*/

extern     void    cr2dif(int source_addr, int fft_length ,int log2_fft_length,int dest_addr);
extern    void    accumulate( int source_addr, int dest_addr , int fft_size );
extern    void    window(int source ,int destination ,int fft_size);
extern    void    Inc_BR ();

void      Process_data_block (NextRead,FftSize,LogFft,AccuStore)
{
          window(NextRead , NextRead, FftSize);
          cr2dif(NextRead , FftSize , LogFft , NextRead);
          accumulate( NextRead, AccuStore , FftSize);
          Inc_BR ();
}
```

Appendix L

Source code: window.asm

```
*************************************************************************
*                            HANNING WINDOWING FUNCTION
*
*       Function written by Kristian Jorgensen PS/RF CERN  Oktober 1998
*
*************************************************************************
*
*    SYNOPSIS: window = (source_addr , dest_addr ,  fft_size )
*    is loaded from stack to:  AR0         AR2            R2
*
*       int       *source_addr : the address of the interleaved complex source input
*       int       *dest_addr   : the destination address
*       int       fft_size     : the length of the fft
*
*    There are two tap tables, one for 256 and one for 512 ffts.
*    The program itself finds out what table to use, but they both
*    have to be linked with the rest of the program. The tap table
*    files are called:
*
*               TAPS256.asm and TAPS512.asm
*
*    Currently they are Hanning tap coefficients generated in MATLAB.
*
*    The signal is interleaved complex data which is overwritten by the
*    windowed version, so the source address is the same as the destination
*    address. The input data transfered from FIFO are 16 bit 2's complement
*    integer values situated in the upper MSBs of the 32 bit word.
*    This function is correcting the 16 LSBs to 0000h or FFFFh depending
*    on the sign.
*
*    Parameters are passed by the stack, ONLY by the stack, remeber to include the
*    library RTS40.LIB .
*
*************************************************************************
        .globl  _tap256
        .globl  _tap512
        .globl  _window


         .sect   ".data"
WIN256   .word   _tap256
WIN512   .word   _tap512
POS_COR  .word   11111111111111110000000000000000b    ; bug: FFFF0000h not possible ?
NEG_COR  .word   0000FFFFh
```

```
        .sect   ".text"
_window
        LDI     SP,AR0
        PUSH    DP
        PUSH    R0
        PUSH    R1
        PUSH    R2
        PUSH    R3
        PUSH    R4
        PUSH    R5
        PUSH    R6
        PUSHF   R6
        PUSH    AR1
        PUSH    AR2
        PUSH    AR3
        PUSH    AR4
        PUSH    AR5
        PUSH    AR6
        PUSH    R8
* stack parameter passing
        LDI     *-AR0(1),AR5            ; Next read cell, centent is source address
        LDI     *-AR0(2),AR3            ; Next store cell, content is destination address
        LDI     *-AR0(3),AR1            ; FftSize cell, content FFT_size


        LDI     *AR5,AR4                ; AR4 = pointer to source address
        LDI     *AR3,AR2                ; AR2 = pointer to destination address
        LDI     *AR1,R2
        LDI     R2,RC                   ; set repetition counter
        LDP     WIN256                  ; set data-page pointer
        LDI     @WIN256,AR1             ; default, AR1 start at 256-tap-table
        LDI     @POS_COR,R0             ; correction of 16 LSBs for positive number
        LDI     @NEG_COR,R1             ; correction of 16 LSBs for negative number
        CMPI    100h,RC                 ; is default right ?
        BZ      DEFAULT                 ; branch if default approved
        LDI     @WIN512,AR1             ; new tap-length, AR1 start at 512-tap-table
DEFAULT:LDI     02h,IR1


        RPTB    BLK
* correct 16 bit number to right 32 bit represenation
* real part -> R3
        CMPI    0h,*AR4
        BN      NEGR                    ; if the number pointed to is negative, branch
        AND     R0,*AR4,R3              ; corrected 16 bit integer number stored in R3
        FLOAT   R3                      ; convert integer value to float
        BU      IMAG
NEGR:   OR      R1,*AR4,R3              ; corrected 16 bit integer number stored in R3
        FLOAT   R3                      ; convert integer value to float
* imaginary part -> R4
IMAG:   CMPI    0h,*+AR4
        BN      NEGI                    ; if the number pointed to is negative, branch
        AND     R0,*+AR4,R4             ; corrected 16 bit integer number stored in R3
        FLOAT   R4                      ; convert integer value to float
        BU      TAP
NEGI:   OR      R1,*+AR4,R4             ; corrected 16 bit integer number stored in R3
        FLOAT   R4                      ; convert integer value to float
* multiplying with tap
```

```
TAP:    MPYF    *AR1,R3                 ; real * tap = R3
        STF     R3,*AR4++               ; R3 stored
        MPYF    *AR1++,R4               ; real * tap = R4
BLK:    STF     R4,*AR4++               ; R4 stored

        POP     R8
        POP     AR6                     ; Restore the register values and return
        POP     AR5
        POP     AR4
        POP     AR3
        POP     AR2
        POP     AR1
        POPF    R6
        POP     R6
        POP     R5
        POP     R4
        POP     R3
        POP     R2
        POP     R1
        POP     R0
        POP     DP
        RETS
        .end
```

Appendix M

Source code: cr2dif.asm

```
*******************************************************************************
*                        FFT ALGORITHM
*
*       FUNCTION WRITTEN/MODIFIED BY KRISTIAN JORGENSEN PS/RF CERN  OKTOBER 1998
*
*******************************************************************************
*
*  FILENAME     : CR2DIF.ASM
*
*  DESCRIPTION  : COMPLEX, RADIX-2 DIF FFT FOR TMS320C40 (C CALLABLE)
*
*  DATE         : 6/93
*
*  VERSION      : 4.0
*
*******************************************************************************
*
*  VERSION        DATE         COMMENTS
*  -------        ----         --------
*   1.0           10/87        PANNOS PAPAMICHALIS (TI HOUSTON)
*                              ORIGINAL RELEASE
*   2.0           1/91         DANIEL CHEN (TI HOUSTON): C40 PORTING
*   3.0           7/91         ROSEMARIE PIEDRA (TI HOUSTON): MADE IT C-CALLABLE
*   4.0           6/93         ROSEMARIE PIEDRA (TI HOUSTON): ADD SUPPORT FOR
*                              IN-PLACE BIT-REVERSING
*
*******************************************************************************
*
*  SYNOPSIS: INT  CR2DIF(SOURCE_ADDR,FFT_SIZE,LOGFFT,DST_ADDR)
*                              AR2        R2       R3    RC
*
*            FLOAT   *SOURCE_ADDR    ; INPUT ADDRESS
*            INT     FFT_SIZE        ; 64, 128, 256, 512, 1024, ...
*            INT     LOGFFT          ; LOG (BASE 2) OF FFT_SIZE
*            FLOAT   *DST_ADDR       ; DESTINATION ADDRESS
*
*   - THE COMPUTATION IS DONE IN-PLACE.
*   - SECTIONS TO BE ALLOCATED IN LINKER COMMAND FILE: .FFTTXT : FFT CODE
*                                                      .FFTDAT : FFT DATA
*
*******************************************************************************
*
*  DESCRIPTION:
*
```

```
*  GENERIC PROGRAM FOR A RADIX-2 DIF FFT COMPUTATION USING THE TMS320C4X
*  FAMILY. THE COMPUTATION IS DONE IN-PLACE AND THE RESULT IS BIT-REVERSED.
*  THE PROGRAM IS TAKEN FROM THE BURRUS AND PARKS BOOK, P. 111.
*  THE INPUT DATA ARRAY IS 2*FFT_SIZE-LONG WITH REAL AND IMAGINARY DATA
*  IN CONSECUTIVE MEMORY LOCATIONS: RE-IM-RE-IM
*
*  THE TWIDDLE FACTORS ARE SUPPLIED IN A TABLE PUT IN A SECTION
*  WITH A GLOBAL LABEL _SINE POINTING TO THE BEGINNING OF THE TABLE
*  THIS DATA IS INCLUDED IN A SEPARATE FILE TO PRESERVE THE GENERIC
*  NATURE OF THE PROGRAM. THE SINE TABLE SIZE  IS (5*FFT_SIZE)/4.
*
*  NOTE: SECTIONS NEEDED IN THE LINKER COMMAND FILE: .FFTTXT : FFT CODE
*                                                    .FFTDAT : FFT DATA
*
*******************************************************************************
*                          +
*   AR + J AI ------------------------------------- AR' + J AI'
*                     \         / +
*                       \     /
*                         \ /
*                         / \
*                       /     \
*                     /         \ +
*   BR + J BI ----------------------- COS - J SIN ---- BR' + J BI'
*
*
*   AR'= AR + BR
*   AI'= AI + BI
*   BR'= (AR-BR)*COS + (AI-BI)*SIN
*   BI'= (AI-BI)*COS - (AR-BR)*SIN
*
*******************************************************************************
        .GLOBL  _SINE               ; ADDRESS OF SINE/COSINE TABLE
        .GLOBL  _CR2DIF             ; ENTRY POINT FOR EXECUTION
        .GLOBL  STARTB,ENDB         ; STARTING/ENDING POINT FOR BENCHMARKS

        .SECT   ".DATA"
SINTAB  .WORD   _SINE
OUTPUTP .SPACE  1
FFTSIZE .SPACE  1

        .SECT   ".TEXT"
_CR2DIF:
        LDI     SP,AR0
        PUSH    DP
        PUSH    R4                  ; SAVE DEDICATED REGISTERS
        PUSH    R5
        PUSH    R6                  ; LOWER 32 BITS
        PUSHF   R6                  ; UPPER 32 BITS
        PUSH    AR4
        PUSH    AR5
        PUSH    AR6
        PUSH    R8
        LDP     SINTAB

    .IF .REGPARM == 0
        LDI     *-AR0(1),AR4        ; !!  points to cell with data source poiter
        LDI     *AR4,AR2            ; AR2 = data source pointer
        LDI     *-AR0(2),AR1        ; !!  points to cell with fft size
        LDI     *AR1,R1             ; R10=N
        LDI     *-AR0(3),AR6        ; !!  points to cell with LogFft
        LDI     *AR6,R9             ; R9 = holds the remain stage number
```

```
        LDI     *-AR0(4),AR5        ; !!  Points to cell with destination pointer
        LDI     *AR5,RC             ; RC = destination pointer
    .ELSE
        LDI    R2,R10
        LDI    R3,R9
    .ENDIF
        STI     RC, @OUTPUTP
        STI     R10,@FFTSIZE

STARTB:
        LDI     1,R8                ; INITIALIZE REPEAT COUNTER OF FIRST LOOP
        LSH3    1,R10,IR0           ; IR0=2*N1 (BECAUSE OF REAL/IMAG)
        LSH3    -1,R10,IR1          ; !! DEFAULT IR1=N/2, POINTER FOR SIN/COS TABLE
        LDI     2,AR5               ; !! BY DEFAULT IE=2
        CMPI    100H,R10            ; !! DECIDE INITIALIZED STEP INDEX IE
        BZ      DEFAULT             ; !!
        LDI     1,AR5               ; !! INIT IE=1
        LSH3    -2,R10,IR1          ; !! IR1=N/4, POINTER FOR SIN/COS TABLE
DEFAULT:LSH     1,R10               ; !! ONLY LABEL INTRODUCED REST ORIGINAL CODE
        SUBI3   1,R8,RC             ; RC SHOULD BE ONE LESS THAN DESIRED #

*       OUTER LOOP
LOOP:
        RPTBD   BLK1                ; SETUP FOR FIRST LOOP
        LSH     -1,R10              ; N2=N2/2
        LDI     AR2,AR0             ; AR0 POINTS TO X(I)
        ADDI    R10,AR0,AR6         ; AR6 POINTS TO X(L)

*       FIRST LOOP
        ADDF    *AR0,*AR6,R0        ; R0=X(I)+X(L)
        SUBF    *AR6++,*AR0++,R1    ; R1=X(I)-X(L)
        ADDF    *AR6,*AR0,R2        ; R2=Y(I)+Y(L)
        SUBF    *AR6,*AR0,R3        ; R3=Y(I)-Y(L)
        STF     R2,*AR0--           ; Y(I)=R2  AND...
||      STF     R3,*AR6--           ; Y(L)=R3
BLK1    STF     R0,*AR0++(IR0)      ; X(I)=R0  AND...
||      STF     R1,*AR6++(IR0)      ; X(L)=R1 AND AR0,2 = AR0,2 + 2*N
*  IF THIS IS THE LAST STAGE, YOU ARE DONE
        SUBI    1,R9
        BZD     ENDB
*       MAIN INNER LOOP
        LDI     2,AR1               ; INIT LOOP COUNTER FOR INNER LOOP
        LDI     @SINTAB,AR4         ; INITIALIZE IA INDEX (AR4=IA)
        ADDI    AR5,AR4             ; IA=IA+IE; AR4 POINTS TO COSINE !! POINTS TO SINE ?
        ADDI    AR2,AR1,AR0           ; (X(I),Y(I)) POINTER
        SUBI    1,R8,RC             ; RC SHOULD BE ONE LESS THAN DESIRED #
INLOP:
        RPTBD    BLK2               ; SETUP FOR SECOND LOOP
        ADDI    R10,AR0,AR6         ; (X(L),Y(L)) POINTER
        ADDI    2,AR1
        LDF     *AR4,R6             ; R6=SIN

*       SECOND LOOP
        SUBF    *AR6,*AR0,R2        ; R2=X(I)-X(L)
        SUBF    *+AR6,*+AR0,R1        ; R1=Y(I)-Y(L)

        MPYF    R2,R6,R0            ; R0=R2*SIN AND...
||      ADDF    *+AR6,*+AR0,R3        ; R3=Y(I)+Y(L)

        MPYF    *+AR4(IR1),R1,R3    ; !! LINE REPAIRED, AR4 AND R1 SWITCHED ; R3 = R1 * COS AND ...
||      STF     R3,*+AR0(1)         ; Y(I)=Y(I)+Y(L)
```

```
        SUBF    R0,R3,R4            ; R4=R1*COS-R2*SIN

        MPYF    R1,R6,R0            ; R0=R1*SIN AND...
||      ADDF    *AR6,*AR0,R3        ; R3=X(I)+X(L)

        MPYF    *+AR4(IR1),R2,R3    ; !! LINE REPAIRED, AR4 AND R2 SWITCHED : R3 = R2 * COS AND...
||      STF     R3,*AR0++(IR0)      ; X(I)=X(I)+X(L) AND AR0=AR0+2*N1

        ADDF    R0,R3,R5            ; R5=R2*COS+R1*SIN
BLK2    STF     R5,*AR6++(IR0)      ; X(L)=R2*COS+R1*SIN, INCR AR6 AND...
||      STF     R4,*+AR6            ; Y(L)=R1*COS-R2*SIN

        CMPI    R10,AR1
        BNEAF   INLOP               ; LOOP BACK TO THE INNER LOOP
        ADDI    AR5,AR4             ; IA=IA+IE; AR4 POINTS TO COSINE
        ADDI    AR2,AR1,AR0         ; (X(I),Y(I)) POINTER
        SUBI    1,R8,RC

        LSH     1,R8                ; INCREMENT LOOP COUNTER FOR NEXT TIME
        BRD     LOOP                ; NEXT FFT STAGE (DELAYED)
        LSH     1,AR5               ; IE=2*IE
        LDI     R10,IR0             ; N1=N2
        SUBI3   1,R8,RC


ENDB:

*****************************************************************************
*------------ BITREVERSAL ----------------------------------------------*
* THIS BIT-REVERSAL SECTION ASSUME INPUT AND OUTPUT  IN RE-IM-RE-IM FORMAT  *
*****************************************************************************

        CMPI    @OUTPUTP,AR2
        BEQD    INPLACE
        NOP
        LDI     @FFTSIZE,IR0            ; IR0 = FFT_SIZE
        SUBI    2,IR0,RC                ; RC  = FFT_SIZE-2
                                        ; SRC != DST
                                        ; AR2 = SRC_ADDR
        RPTBD   BITRV1
        LDI     2,IR1                   ; IR1 = 2
        LDI     @OUTPUTP,AR1            ; AR1 = DST_ADDR
        LDF     *+AR2(1),R0             ; READ FIRST IM VALUE

        LDF     *AR2++(IR0)B,R1
||      STF     R0,*+AR1(1)
BITRV1  LDF     *+AR2(1),R0
||      STF     R1,*AR1++(IR1)

        BUD     END
        LDF     *AR2++(IR0)B,R1
||      STF     R0,*+AR1(1)
        NOP
        STF     R1,*AR1


INPLACE
        RPTBD   BITRV2                  ;
        LDI     AR2,AR1                 ; AR1 = AR2 = SRC_ADDR = DST_ADDR
        NOP     *++AR1(2)
        NOP     *AR2++(IR0)B
```

```
        CMPI    AR1,AR2
        BGEAT   CONT
        LDF     *AR1,R0
||      LDF     *AR2,R1
        STF     R0,*AR2
||      STF     R1,*AR1
        LDF     *+AR1(1),R0
||      LDF     *+AR2(1),R1

        STF     R0,*+AR2(1)
||      STF     R1,*+AR1(1)
CONT    NOP     *++AR1(2)
BITRV2  NOP     *AR2++(IR0)B


;
; RETURN TO C ENVIRONMENT.
;

END:    POP     R8
        POP     AR6      ; RESTORE THE REGISTER VALUES AND RETURN
        POP     AR5
        POP     AR4
        POPF    R6
        POP        R6
        POP     R5
        POP     R4
        POP     DP
        RETS
        .END
```

Appendix N

Source code: accu.asm

```
*************************************************************************
*                        SQUARE    AND ACCUMULATE
*
*       Function written by Kristian Jorgensen PS/RF CERN  Oktober 1998
*
*************************************************************************
*
* SYNOPSIS: accumulate( source_addr, dest_addr , fft_length)
* is loaded from stack to:    AR1         AR2         R2
*
*       int     *source_addr: at beginning of source vector
*       int     *dest_addr    : at beginning of destination vector
*       int     fft_length    : 512 or 1024
*
*        z = A + jB   =>   |z|^2 = A^2 + B^2
*
*       RAM0 memory cells:      dst_addr memory cells:
*               A1                      A1^2 - B1^2
*               B1                       .......
*               A2                       .......
*               B2....                   .......
*
*************************************************************************

        .globl  _accumulate
        .sect   ".data"
        .sect   ".text"
_accumulate

        LDI     SP,AR0              ; register used for passing parameters on stack
        PUSH    DP
        PUSH    R4
        PUSH    R5
        PUSH    R6
        PUSHF   R6
        PUSH    AR4
        PUSH    AR5
        PUSH    AR6
        PUSH    AR7
* stack parameter passing
        LDI     *-AR0(1),AR4            ; address of cell with source address pointer
        LDI     *-AR0(2),AR5            ; address of cell with destination address pointer
        LDI     *-AR0(3),AR7            ; address of cell with FFT length loaded into R2
        LDI     *AR7,R2
```

```
        LDI     *AR4,AR1                ; AR1 = source address
        LDI     *AR5,AR2                ; AR2 = destination address
*pre block loop
        LDI     R2,RC
        SUBI    01h,RC                  ; one element done outside loop, so RC:=RC-1
        LDI     2,IR0                   ; step register of 2

        MPYF    *+AR1,*+AR1,R0          ; R0 = B(1)^2
        MPYF    *AR1,*AR1++(IR0),R1     ; R1 = A(1)^2

* block loop
        RPTB    BLK
        MPYF3   *+AR1,*+AR1,R0          ; R0 = B(x)^2
||      ADDF3   R1,R0,R2                ; R2 = A(x-1)^2 + B(x-1)^2
        MPYF3   *AR1,*AR1++(IR0),R1     ; R1 = A(x)^2
        ADDF    *AR2,R2                 ; mem_cell_value = old_value + new_value
BLK     STF     R2,*AR2++               ; A(x-1)^2 + B(x-1)^2 -> *AR2
Variable syntax.
        POP     AR7
        POP     AR6                     ; Restore the register values and return
        POP     AR5
        POP     AR4
        POPF    R6
        POP     R6
        POP     R5
        POP     R4
        POP     DP
        RETS
        .end
```

Appendix O

Source code: incBR.asm

```
*************************************************************************
*                      MODIFYING BLOCKS READ CELL
*
*       Function written by Kristian Jorgensen PS/RF CERN  November 1998
*
*************************************************************************
*
*    This module is simply counting up the BlocksRead memory cell
*
*************************************************************************
        .globl  _Inc_BR
        .globl  _BlocksRead
        .sect   ".data"
BR      .word   _BlocksRead
        .sect   ".text"
_Inc_BR
        PUSH    AR1
        PUSH    AR5
        PUSH    R0
        PUSH    DP

        LDP     @BR

        LDI     @BR,AR5
        LDI     *AR5,AR1
        LDI     *AR1,R0
        ADDI    01h,R0                  ; add 1 to Block Read
        STI     R0,*AR1                 ; R0 is also return value to c environment

        POP        DP
        POP        R0
        POP        AR5
        POP        AR1
        RETS
        .end
```

Appendix P

Source code: dis_dma.asm

```
*******************************************************************************
*                        DISABLING DMA INTERRUPTS
*
*       Function written by Kristian Jorgensen PS/RF CERN  November 1998

*******************************************************************************
*
*    This module changes the IIF flag for the DMA0 channel, it is set to 0h
*    the rest of the flags are left unchanged. The setting means that the
*    CPU does nolonger receive external interrupts when the FIFO becomes half full.
*    See page 3-13 for IIF flag and 7-17 for interrupt usage in the
*    TMS320C40 User's manual
*
*    Registers used by module R1,R2,R3,IFF, not affected by call
*
*******************************************************************************

        .globl  _disable_dma_int
        .globl  _EIIOFxOff
        .sect   ".data"
F0_mask .word   _EIIOFxOff



        .sect        ".text"

_disable_dma_int
        PUSH    AR1
        PUSH    AR2
        PUSH    R1
        PUSHF   R1
        PUSH    R2
        PUSHF   R2
        PUSH    R3
        PUSHF   R3

* IIF register modification
        LDP     @F0_mask
        LDI     @F0_mask,AR1
        LDI     *AR1,AR2
        LDI     *AR2,R1
        LDI     IIF,R3
        AND     R1,R3                       ; Erase old F0 bits
```

```
        LDI     R3,IIF                      ; set IIF register

        POPF    R3
        POP     R3
        POPF    R2
        POP     R2
        POPF    R1
        POP     R1
        POP     AR2
        POP     AR1
        RETS
        .end
```

Appendix Q

Project timetable

Table Q.1: Task time distribution

	J	F	M	A	M	J	J	A	S	O	N	D	J	F	M	A
Reading	•	•	•	•	•	•										
Syscomm98			•													
C40 Course						•										
Specifications						•	•									
Purchasing						•	•	•								
Learn software							•	•		•						
Write Code						•	•		•	•	•	•	•	•	•	•
Laboratory										•	•	•	•	•		
Real Beam meas.												•				•
Comm. software													•	•		
report			•	•	•				•	•	•					
presentation												•				